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The Proceedings

OF

THE INSTITUTION OF ELECTRICAL ENGINEERS

FOUNDED 1871: INCORPORATED BY ROYAL CHARTER 1921

PART B

ELECTRONIC AND COMMUNICATION ENGINEERING
(INCLUDING RADIO ENGINEERING)

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THE WASHING STAGE IN SEMI-CONDUCTOR MANUFACTURE

Issued by ELGA PRODUCTS LIMITED

The analysis below represents intrinsic* water, considered essential for rinsing transistors and similar devices. There are three methods of producing this very high purity water. Whatever method is used, effluent quality must conform to the analysis.

* intrinsic: belonging naturally—inherent—essential

METHOD 1

Individual line stations—mixed bed column

Observations: Effluent is conductivity water approaching 10 megohm-cm, no regeneration *in situ*. Appears to be satisfactory for most applications.

Limitation: Rather uneconomical in hard water areas.

METHOD 2

Central plant with regeneration in situ, connected to end purification stations

Observations: Effluent quality is "intrinsic water" of the order of 16 megohm-cm+. This approach appears to be generally acceptable and is very economical.

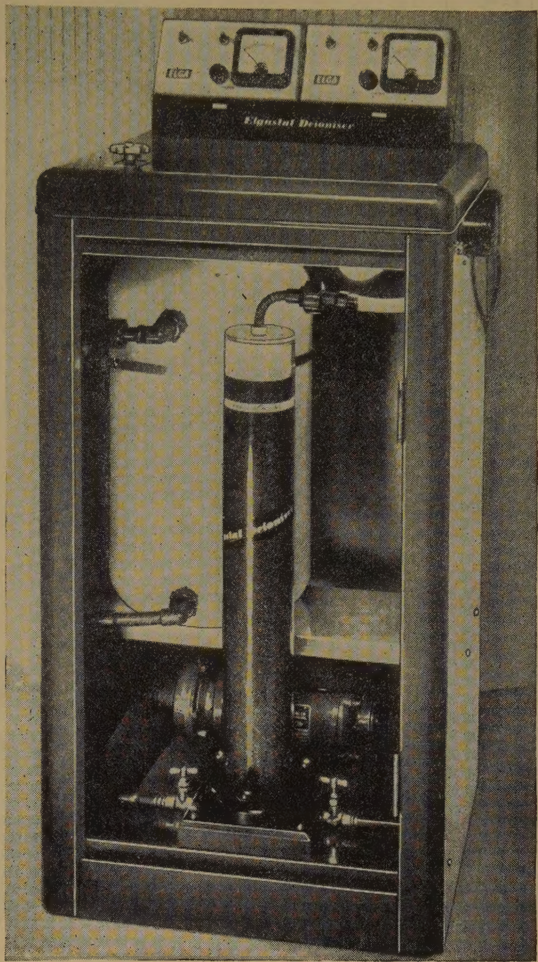
METHOD 3

Individual line stations—mixed bed recirculation

This method produces intrinsic water of the order of 16 megohm-cm+. No regeneration *in situ* or central plant required.

Observations: Acceptable effluent for all known applications, virtually no maintenance. Negligible effluent cost: £2. 10. 0. for 6 to 10 working days. An example of a mixed bed recirculator is shown. This model cost £187.

The Company specialises in the design and manufacture of deionisers for the provision of intrinsic water. Publication PRO/1 includes a paper "Intrinsic Water for Semi-conductor Washing". Please ask for a copy.



The Elgastat Robot B.106.

DEIONISED EFFLUENT ANALYSIS	RESULTS
Trace metals	*Not present*
Silica	*Below 0.1 p.p.m.*
Chloride	*Not present*
Sulphate	*Not present*
Ammonia	*No coloration with Nessler's reagent*
Carbon dioxide	*Not present*
pH value	*6.6-7*
Electrical resistance	*10 megohm-cm+*

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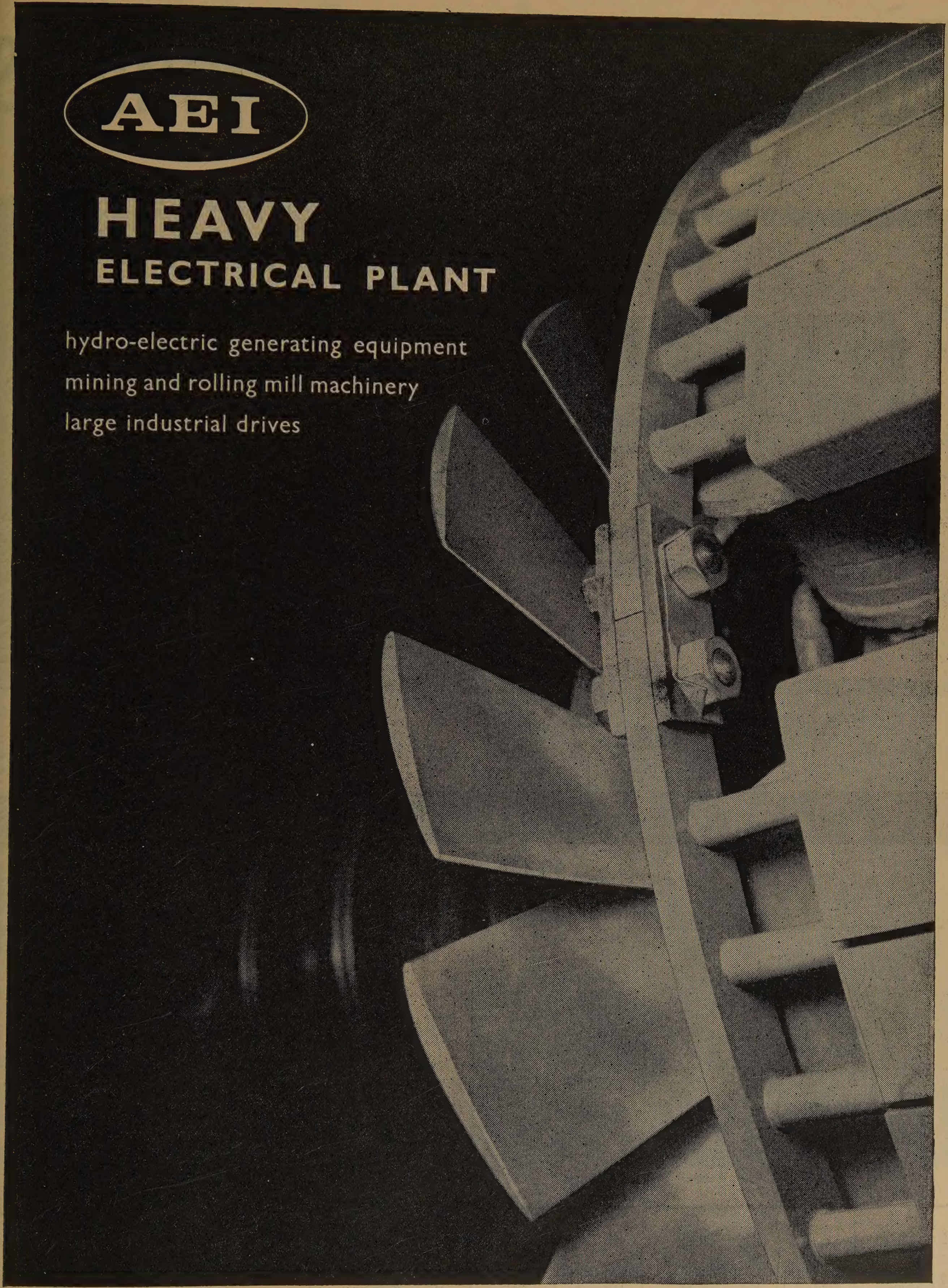

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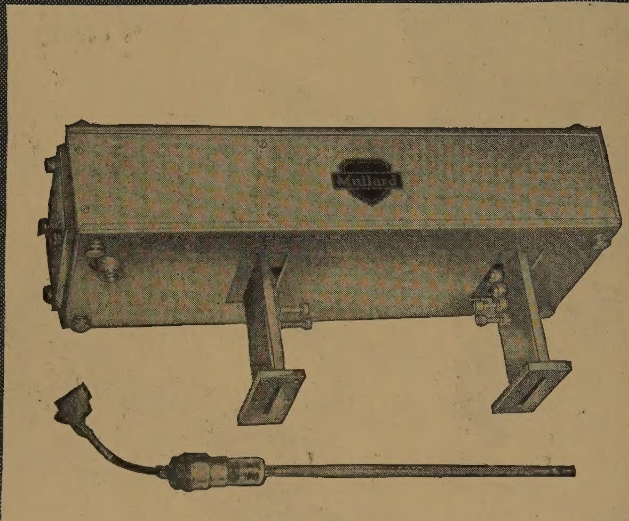
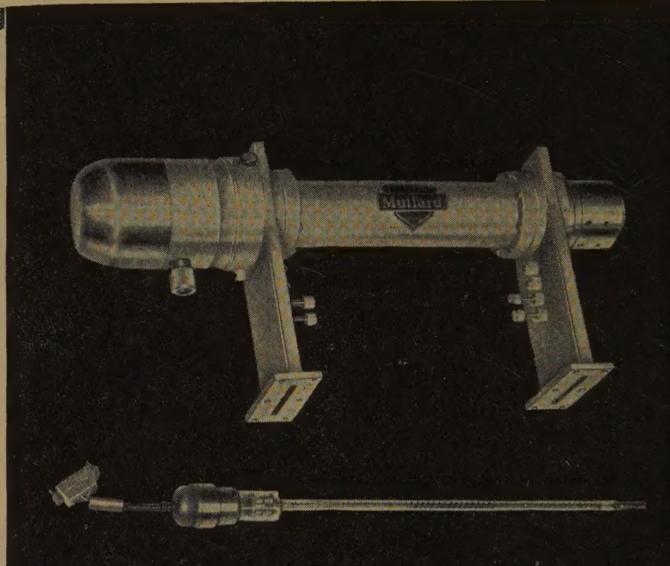


LA4—2



LA4—250

	Tube Type No.	Mount Type No.	Description	Frequency Range (Gc/S)	Power Output (Sat.) (W)	TYPICAL OPERATION AS AMPLIFIER					
						Noise Factor (db)	Power Gain (db)	Attenuation at Ik=0 (db)	Helix Voltage (V)	Collector Voltage (V)	Collector Current (mA)
MICROWAVE LINK TUBES	LA4—2	P4L—1	Low noise tube	3.6 to 4.2	0.002	8.0	27	45	550	500	0.2
	LA4—250	P4L—2	High gain amplifier tube	3.6 to 4.2	0.2	24	31	50	1200	1250	3.5
	LB4—2	P4L—3	Power amplifier tube	3.6 to 4.2	2.5	29	31	60	1300	1250	25
	LB4—8	P4L—4	Power amplifier tube	3.8 to 4.2	8.0	30	37	60	1100	1150	50
	LA6—3	P6L—1	Low noise tube	5.9 to 7.6	0.003	8.0	28	45	600	550	0.3
	LA6—200	P6L—2	High gain amplifier tube	5.9 to 7.6	0.2	25	35	55	1350	1300	5.0
	LB6—12	P6L—3	Power amplifier tube	5.9 to 7.6	12	29	36	65	2400	1500	40
RADAR TUBES	LA9—3	P9L—1 S9L—1 *	High gain broad band tube	7 to 11.5	0.006	23 18 *	27	40	1250	1350	0.55
	LA16—2	P16L—1 S16L—1 *	High gain broad band tube	11.5 to 18	0.003	26 20 *	22	40	1600	1700	0.35

* Solenoid focusing system

with focusing systems

The Mullard range of forward wave amplifier tubes now comprises nine types for microwave link and radar applications.

All tubes in the range are normally supplied with permanent magnet focusing systems and waveguide outputs, but solenoid systems are available on request. The permanent magnet systems eliminate the need for additional power supplies and cooling which is associated with solenoid focusing systems. They are comparatively small and light, and their simplicity makes a significant contribution to equipment reliability and economy.

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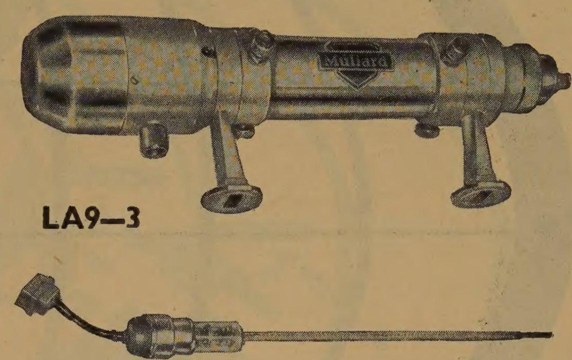
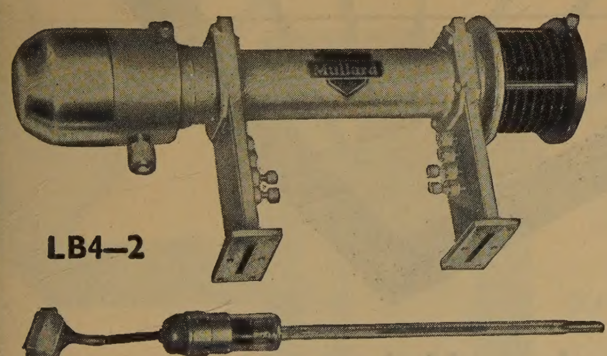
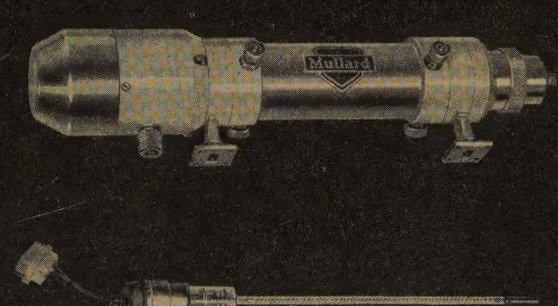
Full details of Mullard travelling wave tubes, their focusing systems and other microwave valves are available from the address below.



LA9—3



LB4—2



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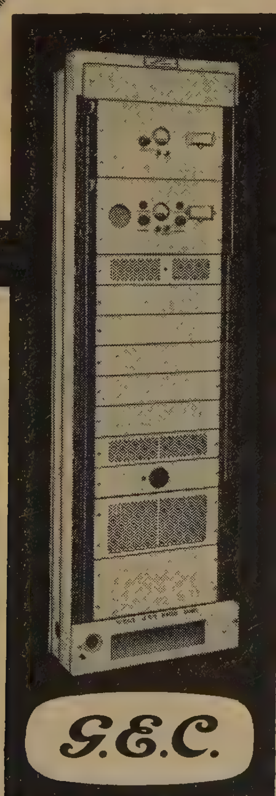

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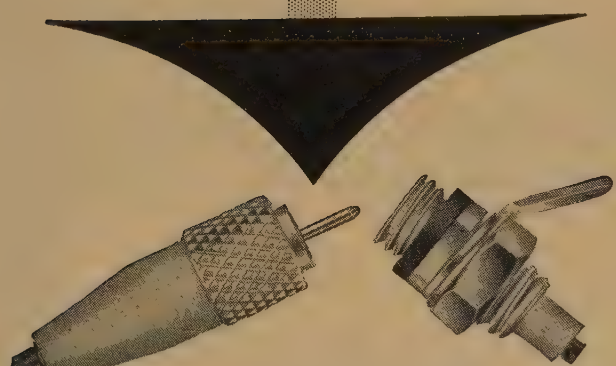

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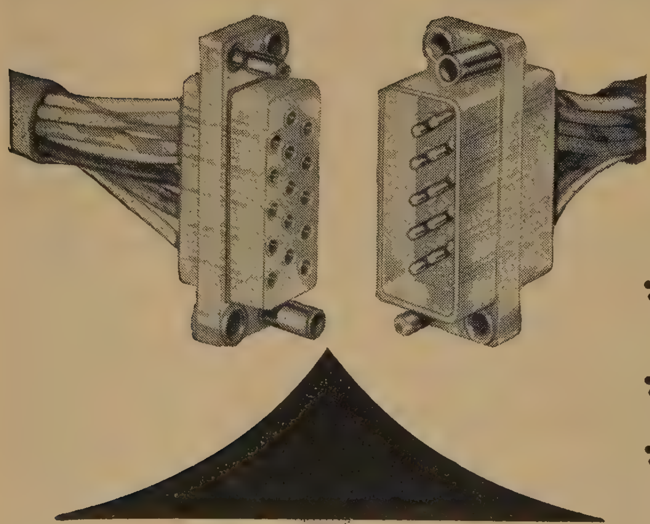

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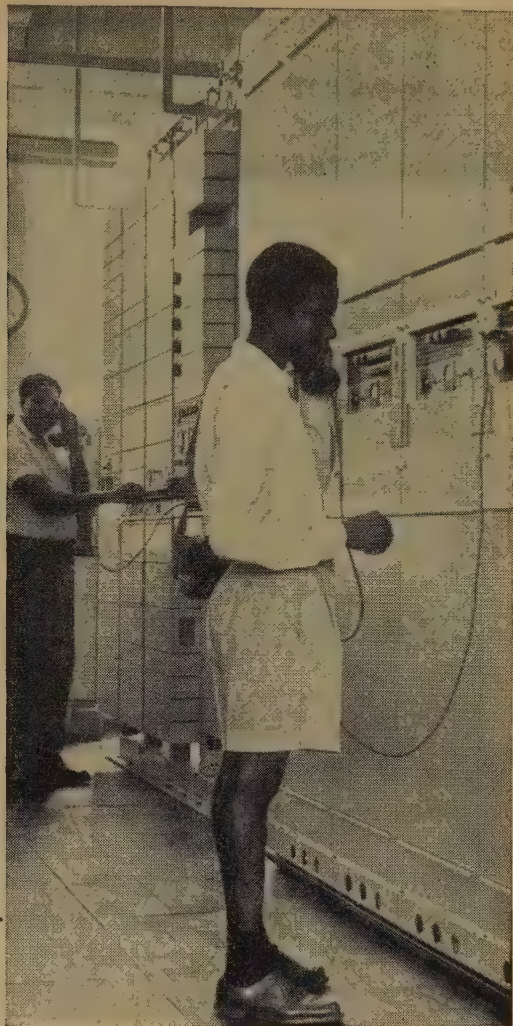
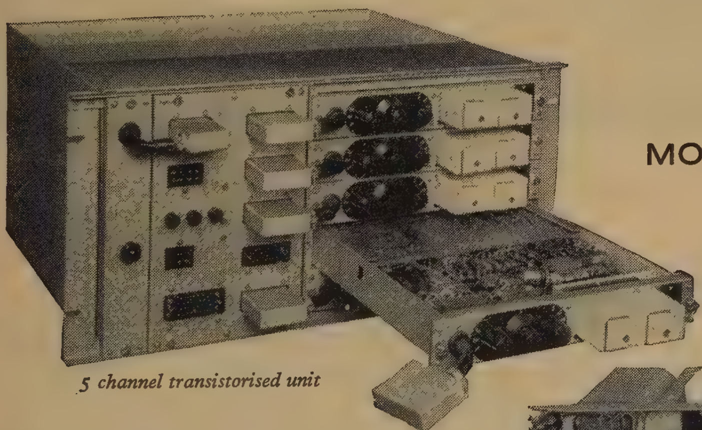


Typical overseas terminal



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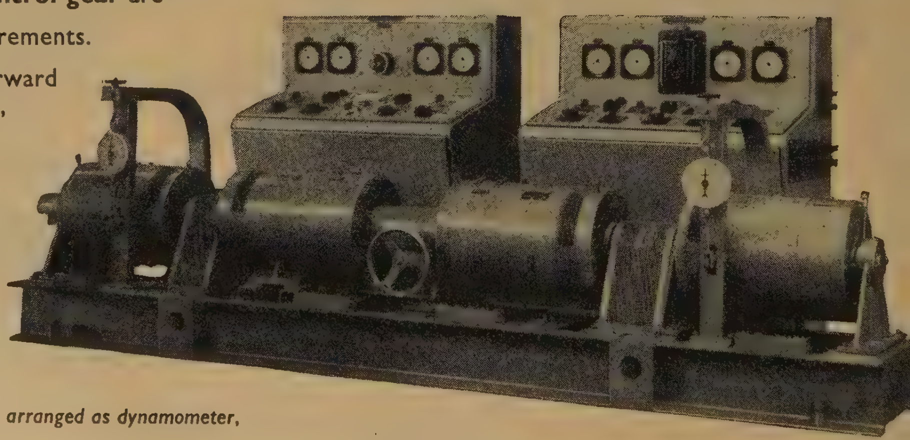

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- *4kVA alternator*
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- *The complete unit can be split into two motor-alternator sets*

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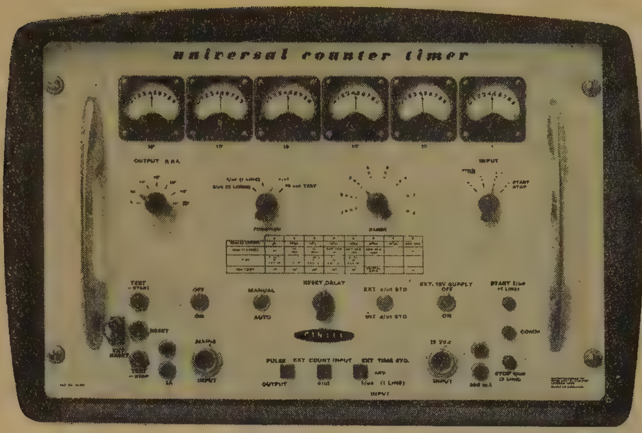
P.28

Transistorized UNIVERSAL COUNTER TIMER



Frequency Measurement

Random Counting

Frequency Division

Time Measurement

Frequency Standard

This fully transistorized portable equipment provides for a wide range of time and frequency measurement as well as facilities for counting, frequency division and the provision of standard frequencies. The facilities available are briefly listed below:

TIME/UNIT EVENT (1 LINE): For the measurement of the time interval between two occurrences in a continuously varying electrical function in the range 3μsec to 1 sec. The time for 1, 10 or 100 such events can be measured.

TIME/UNIT EVENT (2 LINE): For time measurement in range 1μsec to 2777hrs. of any interval defined by a positive or negative going pulse in any combination.

EVENTS/UNIT TIME: For frequency measurement in range 30c/s to 1 Mc/s over period of 0·001, 0·01, 0·1, 1 or 10 secs. Crystal accuracy ± 2 parts in 10^6/week. For mains or 12Vd.c. operation.

Full technical specification available on request.

CINTEL

RANK CINTEL LIMITED
WORSLEY BRIDGE ROAD · LONDON · SE26
HITHER GREEN 4600

Sales and Servicing Agents: Atkins, Robertson & Whiteford Ltd. Industrial Estate, Thornliebank, Glasgow; McKellen Automation Ltd., 122 Seymour Grove, Old Trafford, Manchester, 16; Hawnt & Co. Ltd., 59 Moor St., Birmingham, 4.

right on the button!

with the

PLAN-ETELPHONE

in colour

TWO-TONE GREEN
TWO-TONE GREY
CONCORD BLUE
TOPAZ YELLOW
LACQUER RED
BLACK
IVORY

The Plan-Etelphone can be supplied in FULL TROPICAL FINISH and effectively sealed to prevent the entry of dust and insects.

The Plan-Etelphone, an instrument fulfilling all requirements for plan number working, is the ingenious adaptation of a form already approved by the Council of Industrial Design, accepted by the B.P.O. and now in world-wide service. The extra equipment needed to cater for extension systems, has been skilfully embodied in the available space of the Etelphone without detracting from its aesthetic qualities. Several types are available and when used in conjunction with the Etelphone offer a wide range of extension working facilities. These instruments are obtainable in the same attractive colour range as the Etelphone and can be converted for wall mounting if required.

The instrument illustrated on the opposite page has provision for up to 7 press buttons and up to 3 lamp signals, with a total of buttons and signals not exceeding 7. The main switching, with interlocking and tripping facilities, is provided by a maximum of 4 press buttons mounted immediately in front of the handset and 2 lamp signals or press buttons can be fitted above and on either side of the dial number ring. For further requirements a button or lamp signal is mounted in the dial lock fitting on the front of the case. The mechanical arrangement is completely flexible and will provide any combination of interlocking with or without switch-hook release. Transparent protective covers are used on the springset and very simple assemblies are provided for the less elaborate plans.

The various types of Plan-Etelphone will cater for all existing B.P.O. plan numbers together with E.T. Secretarial Schemes and systems already supplied to other countries. The same instrument is suitable for providing a 1 + 3 House Exchange System based either on the use of multiple cabling or a central relay set. A transistorised ringing unit fitted in the Plan-Etelphone replaces the hand generator previously required on some plan numbers. Power is taken from a 6-volt local battery.

CONVERTIBLE TO WALL MOUNTING

For further information please write to :—

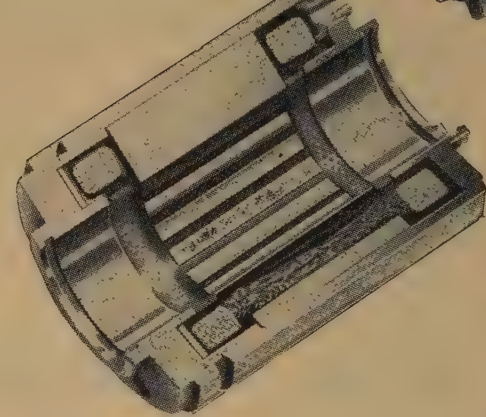
ERICSSON TELEPHONES LIMITED—ETELCO LIMITED

HEAD OFFICE :- 22, LINCOLN'S INN FIELDS, LONDON, W.C.2. TEL : HOL. 6936

WORKS :- BEESTON, NOTTINGHAM AND SUNDERLAND.

Araldite IN SERVOCOMPONENTS

The Synchro units shown, made by R. B. Pullin & Co. Ltd., are sectioned to show how the stators are integrally cast in Araldite to provide maximum protection against the effects of extremes of temperature, humidity and vibration. The excellent machining properties of Araldite make possible a straight-through bore technique, which eliminates errors in alignment and also permits the use of the smallest possible air gap between rotor and stator. High insulation and dielectric strength, remarkable adhesion to metals, and negligible shrinkage on curing make Araldite eminently suitable for use in the construction of precision electrical equipment.



Araldite epoxy resins are used—

- **for casting high grade solid electrical insulation**
- **for impregnating, potting or sealing electrical windings and components**
- **for producing glass fibre laminates**
- **for producing patterns, models, jigs and tools**
- **as fillers for sheet metal work**
- **as protective coatings for metal, wood and ceramic surfaces**
- **for bonding metals, ceramics, etc.**

This photograph shows an A.E.W. electric oven, capable of maintaining temperatures within close limits, as used by R. B. Pullin & Co. Ltd. for curing the Araldite-filled stators.

Araldite *epoxy resins*

Araldite is a registered trade name

CIBA (A.R.L.) LIMITED

Duxford, Cambridge Telephone: Sawston 2121

AP 462

"...THERE'S NOTHING MUCH IN CAPACITORS DAD!"

Sweet little chap! Remarks like that are *so* endearing—especially to the men at S.T.C. who devote their lives on research into Capacitors. Still . . . it's today's enquiring mind that makes tomorrow's genius. So let the lad's investigations continue. He'll soon discover that there's more to S.T.C. Capacitors than meets the cold chisel.

STANTELAC CAPACITORS
STANTELUM ELECTROLYTIC CAPACITORS
MINIATURE STANTELUM ELECTROLYTIC CAPACITORS
STANTELISED METALLISED PAPER CAPACITORS
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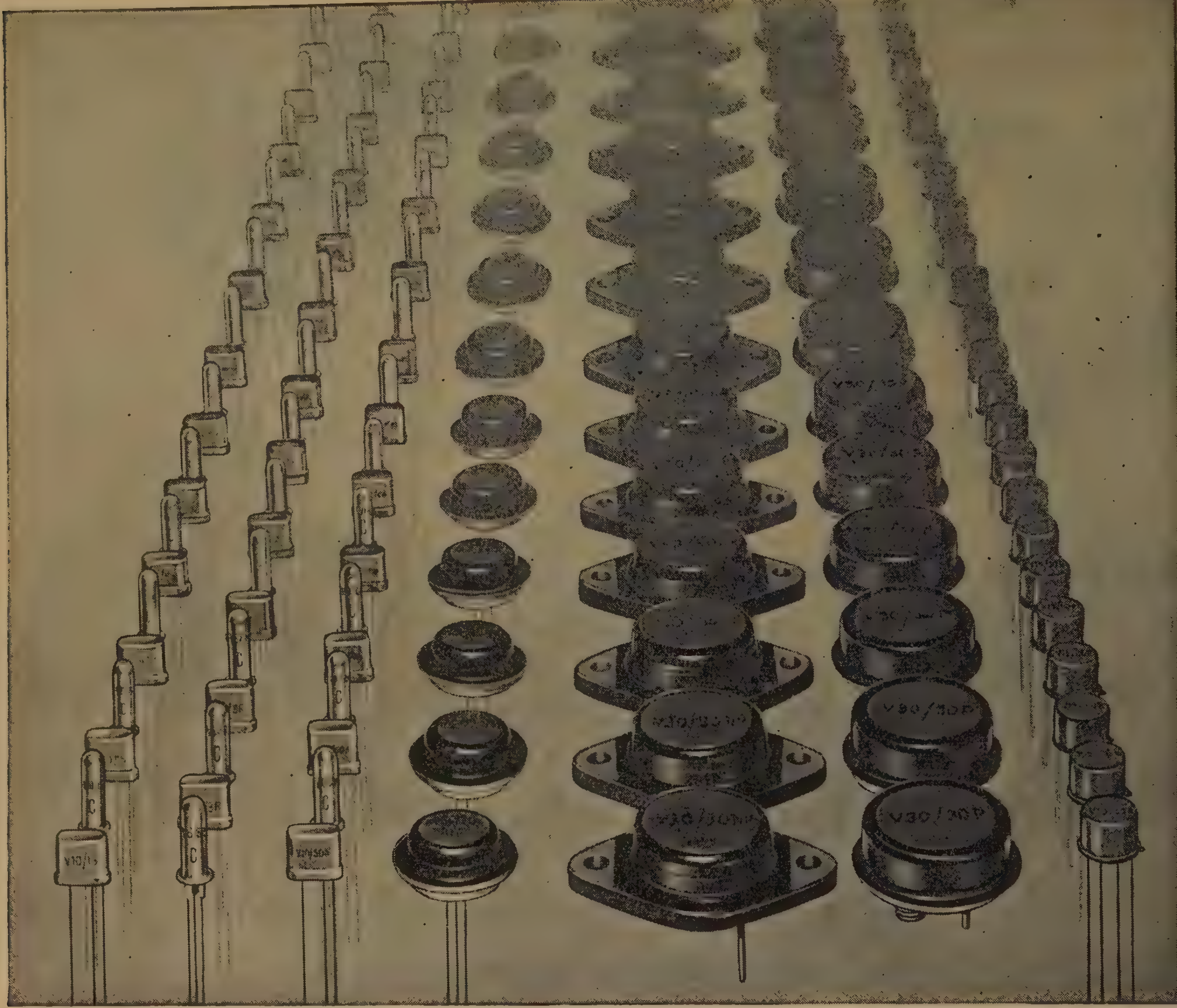
Standard Telephones and Cables Limited
Registered Office: Connaught House, Aldwych, London, W.C.2
CAPACITOR DIVISION : SALES DEPARTMENT · FOOTSCRAY · SIDCUP · KENT



The Newmarket range includes them all

Switching	V10/1S, V10/1SC V10/2S, V10/2SC	Voltage ratings 10, 15, 20V. Switching Rise times down to .1μs (V10/1S), .2μs (V10/2S) Max. dissipation 125mW; Peak current 500mA. Rectangular or K1007/A1/D2 standard cylindrical style can
R.F.	V6/2R, V6/2RC V6/4R, V6/4RC V6/8R, V6/8RC	Voltage ratings 6, 10, 15, 20, 25V. Typ. frequency cut-offs 3, 5.5, 10 Mc/s Max. dissipation 125mW; Rectangular or K1007/A1/D2 standard cylindrical style can
A.F.	V10/15A, V10/15AC V10/30A, V10/30AC V10/50A, V10/50AC	Voltage ratings 10, 15, 30V. Typ. betas 20, 40, 75 Max. dissipation 200mW; Rectangular or K1007/A1/D2 standard cylindrical style can
I.P. (Intermediate Power)	V15/20IP V30/20IP V60/20IP	Voltage ratings 15, 30, 60V. Typ. beta 40 Max. dissipation 2W; Max. current 2Amp.
N.P. (Noodle Power)	V15/15NP | V30/15NP V15/30NP | V30/30NP	Voltage ratings 15, 30V. Typ. betas 25, 40 Max. dissipation 15W; Max. current 6Amp. Standard Diamond (JEDEC E2—42) Base Cold welded case
P. (Power)	V15/10P, V30/10P, V60/10P V15/20P, V30/20P, V60/20P V15/30P, V30/30P, V60/30P	Voltage ratings 15, 30, 60V. Typ. betas 18, 24, 40 Max. dissipation 10W; Max. current 3Amp.
VHF Drift	V15/20R	Voltage rating 15V. Typ. frequency cut-off 30 Mc/s. Max. dissipation 75mW, Max. current 12mA. JEDEC T O-5—welded case.

If you have not received a copy of our booklet "Semi-conductor Device Data", ask us to send you one.

NEWMARKET TRANSISTORS

Newmarket Transistors Ltd

Exning Road, Newmarket, Suffolk Telephone: Newmarket 3381/4 Cables: Semicon Newmark

TA 27

we can get you out of a thyratron problem

We, that's the M-O Valve Co. Ltd., have spent over 40 years bottling up hard valves, soft valves, tough going-aloft valves, standard valves, valves to government specs . . . so we know how you feel in there. But we're used to it, because now we've acquired a rare collective know-how that can beat most valve problems. And of course, it's backed by the brilliant research, progressive development and flexible production you'd expect us to have. Next time you're enveloped in a valve problem, let us know—if we haven't a ready-made type for you, we'll see if we can make you one to measure!

QUADRUPLE-LIFE HYDROGEN THYRATRONS

G.E.C. hydrogen filled thyratrons incorporate a patented titanium hydride replenisher in the heater than maintains the pressure in the valve sensibly constant over a wide temperature and voltage range. This virtually eliminates valve failure due to gas clean up and increases the life expectation of the valve by as much as 300%

Type	Description	v_a (pk) max. (kv)	i_a (pk) max. (A)	I_a (av) max. (A)	* Pulse power × p.r.f. max. (MW-c/s)
GHT2	Directly heated type having small size for its power rating	18	700	1.25	5000
5948	Equivalent to the American 1754/5948 but using the G.E.C. replenisher system to give longer life	25	1000	1	5000
GHT3	Similar to the 5948 but with greatly improved 'jitter' figure (0.001 μsec)	25	1000	1	5000
GHT4	New very high power tube	25	2000	3	10,000

* Defined as $\frac{1}{2} \times v_a(pk) \times i_a(pk) \times$ p.r.f.

G.E.C. valves are obtainable from

THE M-O VALVE CO. LTD

BROOK GREEN · HAMMERSMITH · LONDON W.6

A subsidiary of the General Electric Co. Ltd.

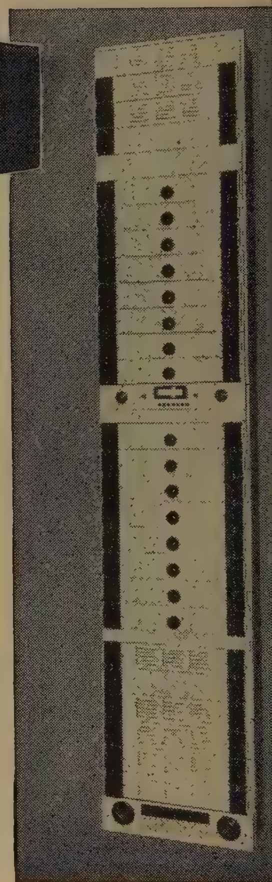
TMC

Telephone and Telegrap

RADIC

CARRIER TELEPHONE SYSTEM

R40C



* 6 kc/s channel spacing
* channel bandwidth 300-3400 c/s
* out-band dialling (E & M) or ringdown facilities
* low-level signalling (-20dbm0) tone-on or tone-off idle
* 40 circuits on one double-sided 9 foot rack including carrier, signalling and power supplies
* alternative arrangements for small systems (up to 16 channels) or for large systems
* conforms with latest CCITT recommendations on levels and impedance for radio equipment

THESE TWO SYSTEMS ARE PART OF THE **T.M.C.** RANGE OF TRANSMISSION EQUIPMENT WHICH INCLUDES:

2, 3, 4 and 6 kc/s spaced carrier telephone equipment for cable, radio and ope wire systems. Transistorized 120 c/s and 170 c/s FM Telegraph Equipment. Transistor VF Repeater Equipment.
Transistorized privacy equipment.

TELEPHONE MANUFACTURI

TRANSMISSION DIVISION · CRAY WO

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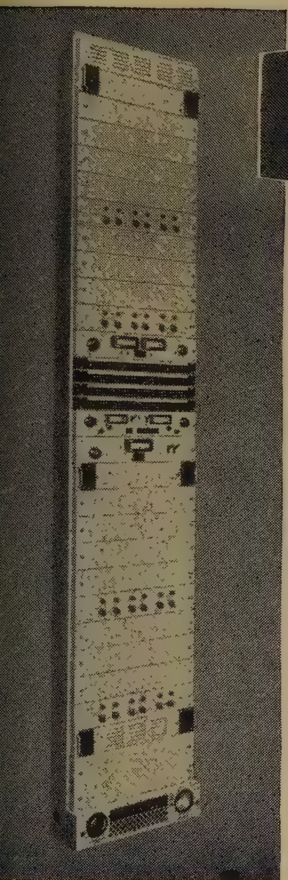

T24D VF TELEGRAPH SYSTEM

- Fully transistorized
- Frequency modulated (±30c/s)
- 120 c/s spaced channels
- 24 bothway channels on one speech circuit (300-3,400 c/s)
- Pilot frequency control, if required
- Plug-in channel units simplify maintenance and extensions
- Robust circuits using only 7 transistors per channel
- Mains or battery operation
- 24 channels with pilot equipment, test panel and double current telegraph supplies on 9′ rack-side

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OTHER COUNTRIES *Transmission Equipment Only)* — Automatic Telephone & Electric Co. Ltd., London.

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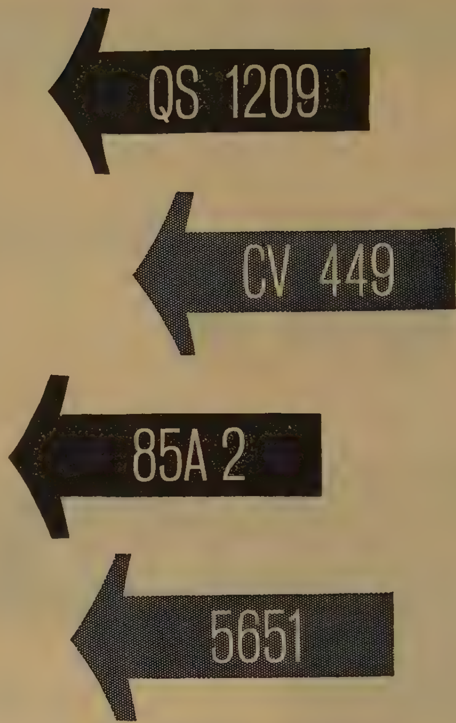


A valve by any other name...

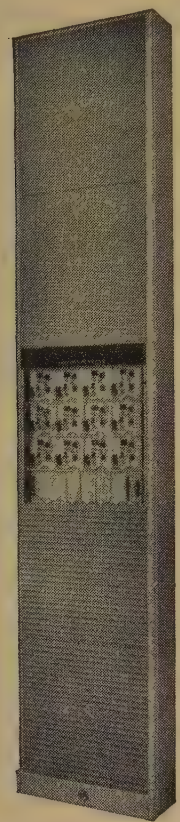
Call it what you will, the **English Electric Valve** voltage stabiliser QS1209 (CV449) is the equivalent of the 5651 and the 85A2, but is superior in performance. It is a cold cathode gas-filled tube with a B7G glass button base, and is one of a large range of voltage stabilisers and reference tubes manufactured by the Company. For complete technical data please write to:—

ENGLISH ELECTRIC VALVE CO. LTD.

Chelmsford, England
Telephone: Chelmsford 3491

AP 128

A.T.E.
12-CHANNEL 2-WIRE CABLE SYSTEM TYPE C12G



- *12 telephone circuits over a single cable pair.*
- *Fully transistorised.*
- *Power-fed and Frogging Repeaters.*
- *C.C.I.T.T. performance.*
- *Inbuilt signalling*

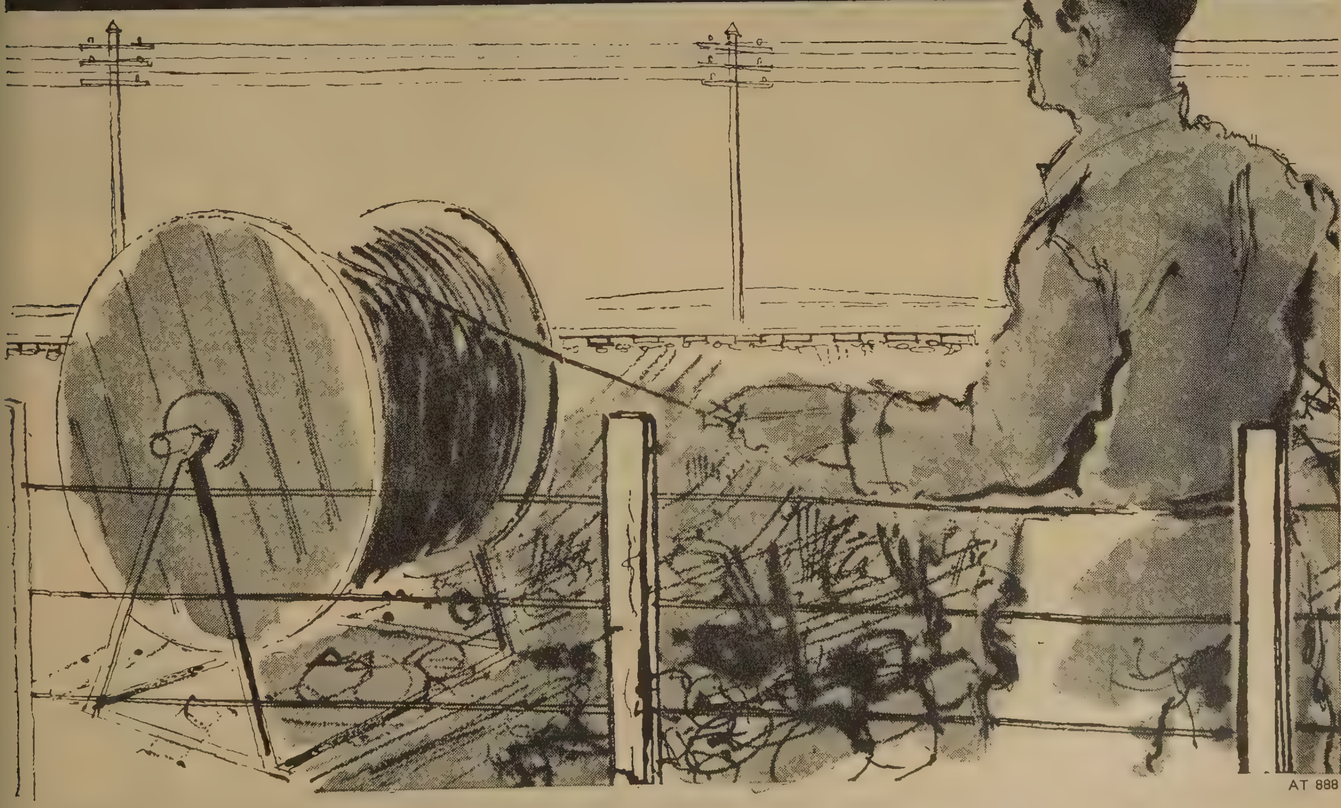
An item in the comprehensive range of A.T.E. Type CM transmission equipment.

For further details write for Leaflet TEL 3201

TRANSMISSION EQUIPMENT TYPE CM.

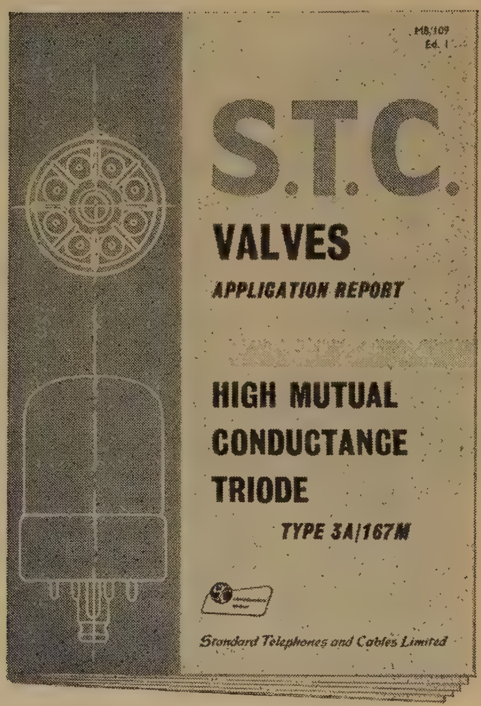
AUTOMATIC TELEPHONE & ELECTRIC CO. LTD

STROWGER HOUSE, ARUNDEL STREET, LONDON, W.C.2.



AT 8881

Application Report for the S.T.C. Type 3A/167M High Mutual Conductance Triode



The 3A/167M is a high performance, long life triode, which has been designed to satisfy the particularly stringent requirements of repeater applications. There is, however, a much wider field of possible applications and this report presents the detailed data necessary to assist the circuit engineer, and the performance which can be expected.

Send for a copy. The number is MS/109.

APPLICATIONS INCLUDE

- WIDE BAND AMPLIFICATION
- LOW NOISE AMPLIFICATION
- V.H.F CASCODE AMPLIFICATION
- PULSE AMPLIFICATION
- LOW DISTORTION AMPLIFICATION
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We are specialists in components. This is no empty phrase: it means that we have been engaged in the design and manufacture of precision-made components for some thirty-five years, and at Enfield we devote a very great deal of our time and space to research, forever seeking new and better methods of production, whilst ensuring that every component maintains our high standard of performance.

Most "Belling-Lee" products are covered by patents or registered designs or applications.

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FUSES · INTERFERENCE FILTERS · RECEIVING AERIALS

MARCONI FM/AM SIGNAL GENERATOR TF 995A/2—*for all-round utility*

The MARCONI Signal Generator Type TF 995A/2 is an accurate and dependable instrument of broadest applicability. It covers from 1·5 to 220 Mc/s in five bands and there are facilities for crystal standardization from 13·5 Mc/s upwards. A precision slow-motion mechanism is employed for the main tuning drive and, for making bandwidth measurements, there is a separate directly-calibrated incremental control. The open-circuit output level is variable, in 1-dB steps, from a minimum of 0·1μV to a maximum of 100 mV at 52 ohms and 200 at 75 ohms. The output may be continuous wave, frequency modulated, amplitude modulated, or simultaneously both frequency and amplitude modulated. The modulation, obtained either from an internal 1000-c/s oscillator or from an external source, is variable to maximum frequency deviations ranging from 25 to 600 kc/s for f.m., and to depths up to 50% for a.m. *Send for leaflet K156 for details.*

Please address enquiries to MARCONI INSTRUMENTS LTD. at your nearest office:

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Telephone: COVent Garden 1234

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23/25 Station Square, Harrogate
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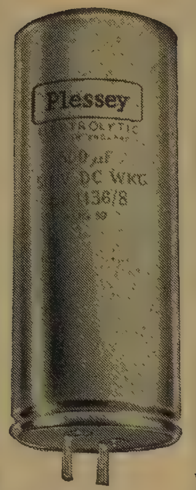
Export Department: Marconi Instruments Ltd., St. Albans, Herts. Telephone: St. Albans 56161

TC 156



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HYPERLYTICS*
give

ten years service

Long shelf life and a working life of 10 years are achieved in the new type of electrolytic capacitor developed by Plessey.

The exacting specifications demanded for specialised electronic equipment have been met by using a high purity etched foil material, and combining the direct Plessey process control with unique manufacturing methods.

With low leakage and high insulation characteristics, Hyperlytics can be supplied in a wide range of values between 0.5 and 12,000 μF. Two types are available, the 1000 series with a temperature range of —30°C to +85°C, and the 11,000 series with a range of —20°C to +50°C.

Capacitance tolerance for both types is normally —20% to +50% but closer limits can be achieved to meet special requirements.

Write for further details of these outstanding new components.

* *High Performance Electrolytic Capacitors*

THE PLESSEY COMPANY LIMITED · CAPACITORS & RESISTORS DIVISION
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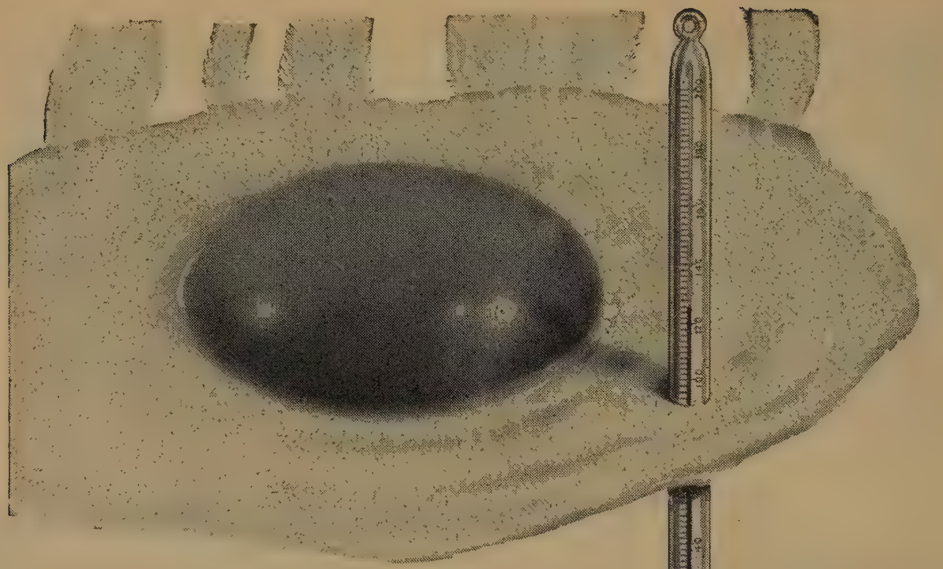
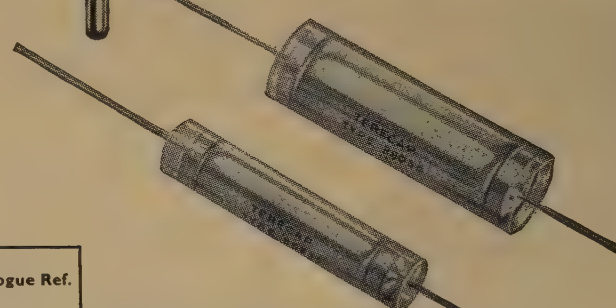
CRI/A



When it's hot enough to fry an egg "*Terecaps*" *are in their element*



SIZES AND RATINGS

Capacitance	Working Voltage d.c. at 71°C	Working Voltage d.c. at 125°C	Test Volts d.c.	Dimensions ins. Length	Dimensions ins. Dia.	Catalogue Ref.
0.1	150	125	300	$1\frac{1}{4}$	$\frac{1}{2}$	8801.C
0.25	150	125	300	$1\frac{7}{8}$	$\frac{1}{2}$	S-8803.C
0.5	150	125	300	$1\frac{3}{8}$	$\frac{5}{8}$	S-8800.C
1.0	150	125	300	$1\frac{7}{8}$	$\frac{3}{4}$	S-8804.C
0.1	250	180	500	$1\frac{1}{4}$	$\frac{1}{2}$	8801.C
0.25	250	180	500	$2\frac{1}{8}$	$\frac{1}{2}$	8803.C
1.0	250	180	500	$2\frac{1}{8}$	$\frac{3}{4}$	8804.C
0.1	350	250	700	$1\frac{5}{8}$	$\frac{1}{2}$	8802.C
0.25	350	250	700	$1\frac{7}{8}$	$\frac{3}{4}$	S-8804.C
1.0	350	250	700	$2\frac{1}{8}$	1	8806.C

Other capacitance values can be supplied to order. We invite your enquiries.

"Terecap"* Capacitors

Dubilier "Terecap" Capacitors are of tubular form with extended foil metal electrodes fitted with wire tail terminations and incorporate a non-hygroscopic film dielectric. Being designed to meet abnormal atmospheric conditions such as obtain in tropical zones the capacitors are supplied hermetically sealed in metal containers with ceramic end-seals.

* *A Registered Dubilier Trade Mark.*

Dubilier "Terecap" Capacitors have these outstanding features:—

1. Can be used up to 125°C. with voltage de-rating above 70°C.
2. High insulation resistance, more than twenty times that of paper dielectric capacitors. (10,000ΩF at 20°C.)
3. Compactness.
4. Excellent capacitance stability over a wide temperature range. (Normal capacitance tolerance ± 20%)
5. Power Factor 0.5% at 20°C. for 1 kc/s.

DUBILIER

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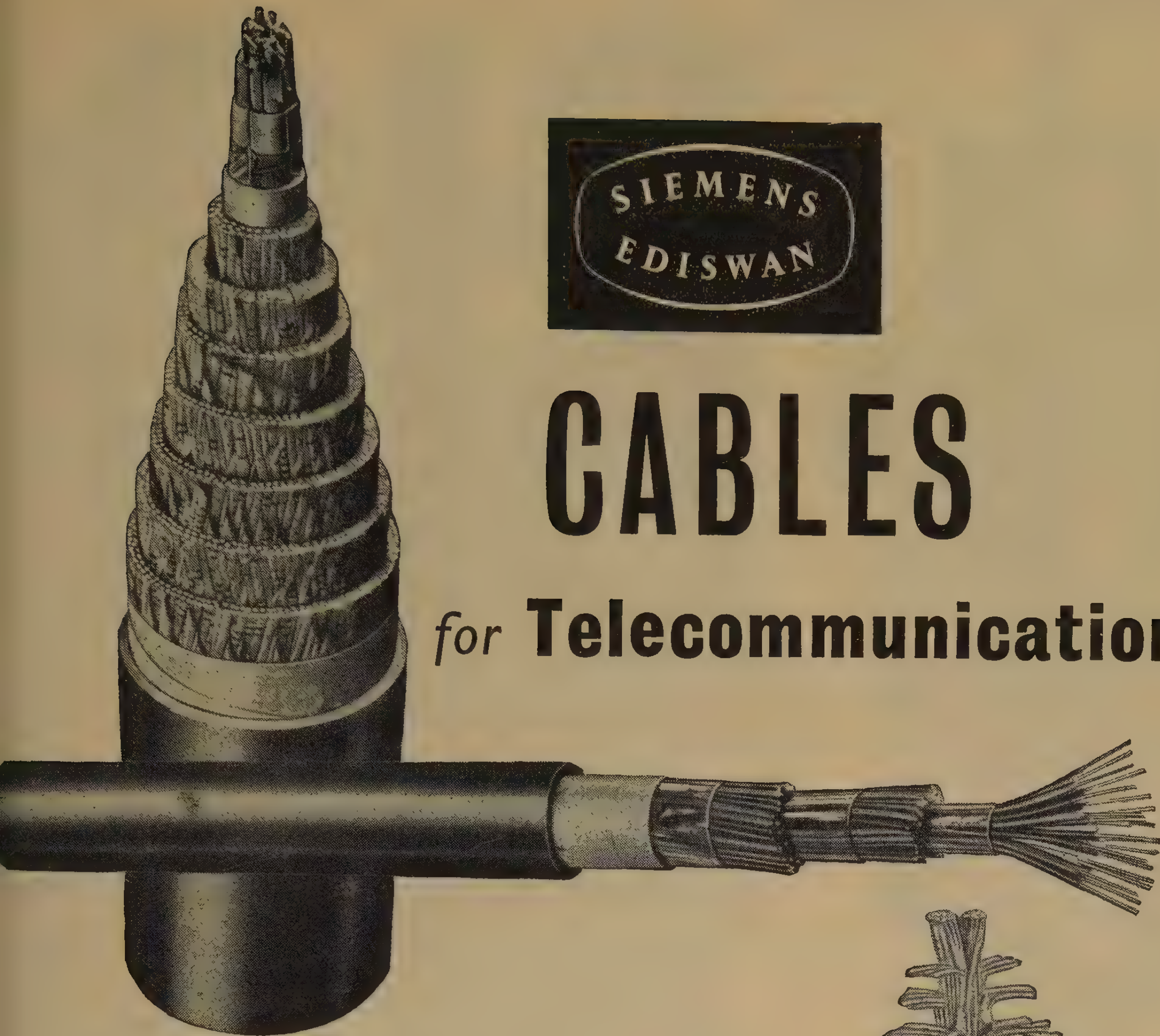
Phone : ACOrn 2241

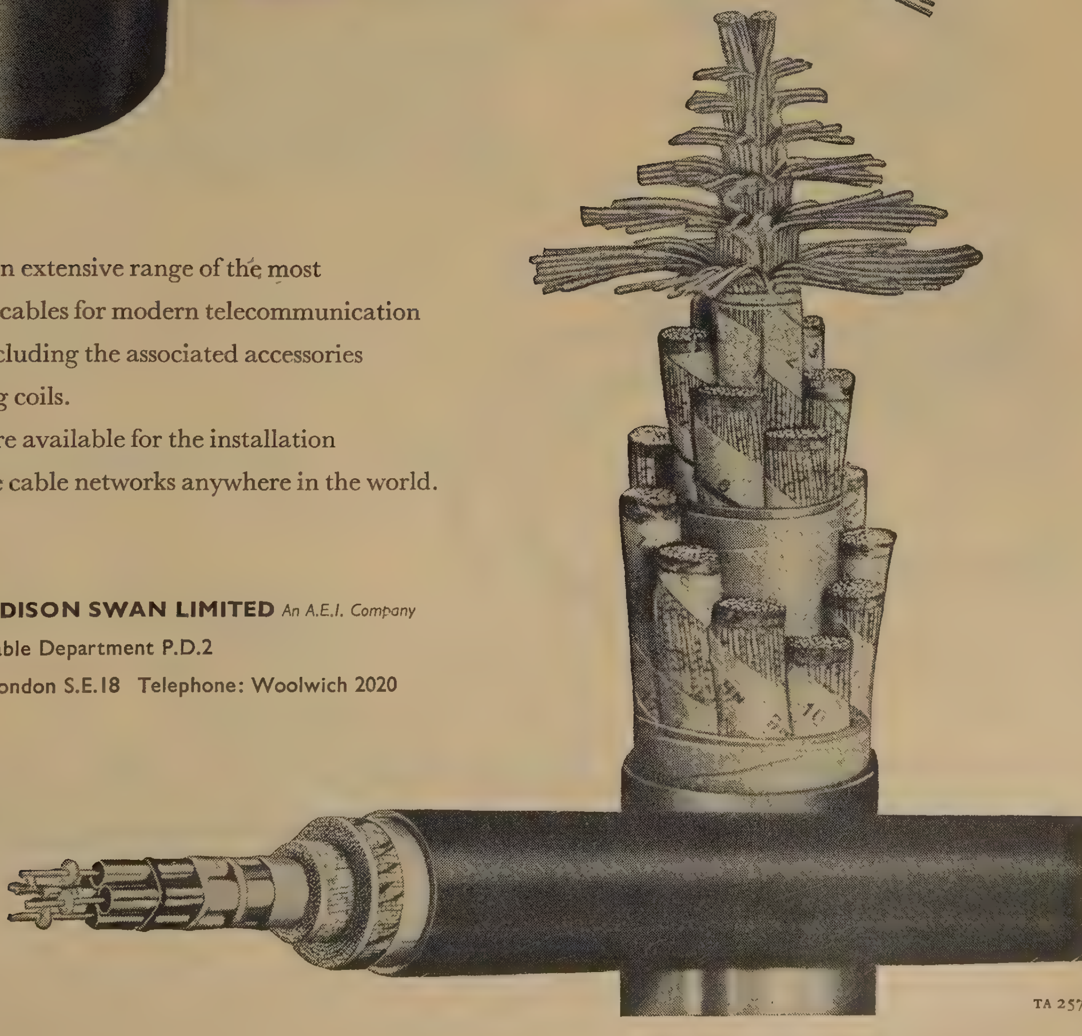
DN.225A

SIEMENS EDISWAN

CABLES

for Telecommunications

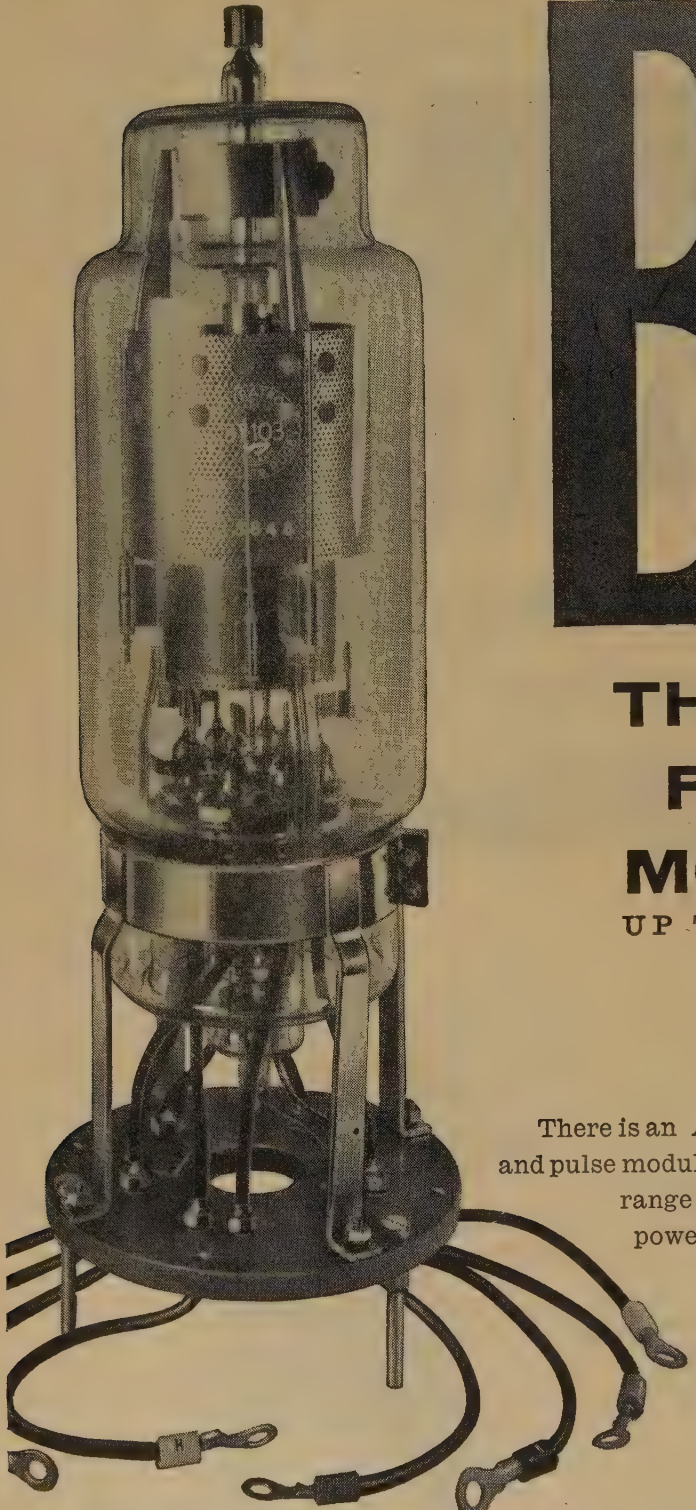




We make an extensive range of the most up-to-date cables for modern telecommunication systems, including the associated accessories and loading coils.
Facilities are available for the installation of complete cable networks anywhere in the world.

SIEMENS EDISON SWAN LIMITED *An A.E.I. Company*
Telephone Cable Department P.D.2
Woolwich, London S.E.18 Telephone: Woolwich 2020

TA 2575



BIG

THYRATRONS FOR PULSE MODULATION

UP TO 1,000 AMPS AT 25kV

BY

AEI

There is an AEI thyratron for every industrial application and pulse modulation requirement. Two recent additions to the range of pulse modulation types, with extremely high power ratings for single and double grid drives, are:—

BT 103

MAX. CATHODE CURRENT 1,000 amp
MAX. ANODE VOLTAGE 25 kV

BT 101

MAX. CATHODE CURRENT 500 amp
MAX. ANODE VOLTAGE 25 kV

Please write for further details and technical data

AEI

Associated Electrical Industries Limited
Electronic Apparatus Division
VALVE & SEMICONDUCTOR SALES
LINCOLN, ENGLAND

A5388

SIEMENS EDISWAN

Cordless Switchboards

Typical of the latest type of 'cordless' telephone switchboards are these operating positions installed by Siemens Edison Swan in one of the British Post Office Trunk Exchanges where operator controlled long distance calls are handled.

These switchboards are noteworthy for their use of remote-controlled automatic selectors in place of outmoded cords and plugs, and for such other advances as automatic call queueing and digit key sending. Calls can be established quickly and without effort, and for the subscriber this means an end to irritating delays.

Leading telephone operating administrations have co-operated in the physical design of the board to ensure the provision of more comfortable and congenial working conditions for the operator, and to eliminate the fatigue associated with conventional type boards.

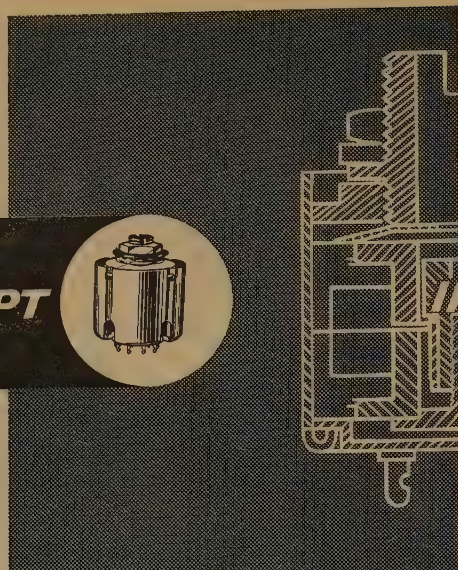
In every important aspect of design, construction and function the Siemens Ediswan cordless switchboard is a further example of our technical leadership in the science of telecommunications gained over many years.

vast experience  **in the science of telecommunications**

SIEMENS EDISON SWAN LIMITED *An A.E.I. Company*
Telecommunications Division P.D.5, Woolwich, London S.E.18. *Telephone: Woolwich 2020*

P.D.5/9

A NEW CONCEPT IN POT CORE DESIGN



Mullard VINKOR range of adjustable pot cores gives you *outstanding* advantages

- Wide range of sizes
- Easily assembled
- Close tolerance permeability
- Precise and easy inductance adjustment
- Stability
- Single hole chassis mounting

Mullard Vinkors are the most efficient adjustable pot core assemblies commercially available. In addition to high performance, they have the distinct advantage of close tolerance permeability, thus enabling designers to precalculate to within ±3% the inductance of the core when wound. Final adjustment, taking into account normal capacitor tolerance, can be easily effected to an accuracy of better than 0.02%, by means of a simple self-locking device built into the core.

Write today for full details of the wide range of Vinkors currently available.

Mullard

VINKOR POT CORES

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COMPUTER TRANSISTORS

The Semiconductors range of Computer Transistors, designed and tested to the special requirements of computer engineers, is the key to a new order of computer speed and reliability. Overall reliability is further increased by making possible a substantial reduction in the number of associated components.

The two types of Silicon Alloy Transistor shortly going into production will make it possible to extend this high-speed computer performance into **ambient** temperatures well above 100°C. **Samples are available now.**

	TYPE	DESCRIPTION	RISE TIME millimicroseconds	Vc max	Ic max
HIGH-SPEED LOW-LEVEL SWITCHING **GERMANIUM**	SB 344 SB 345	General purpose transistors for conventional logic circuits.	50	5v	5mA
	SB 240	Designed for directly coupled circuits. Controlled input, saturation and hole storage characteristics.	30	6v	15mA
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	2N 501	Ultra-high speed transistor with controlled input and saturation characteristics.	10	12v	50mA
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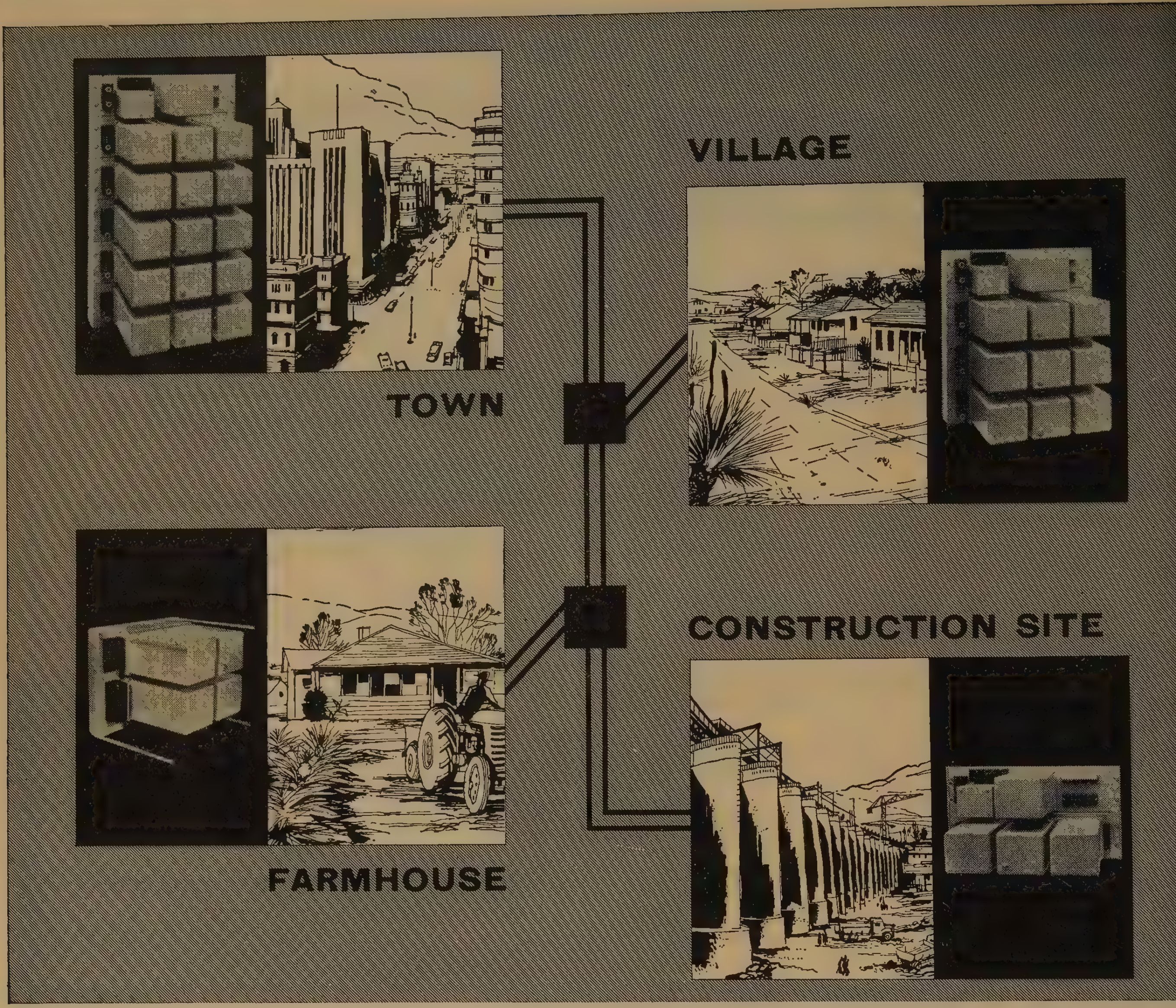

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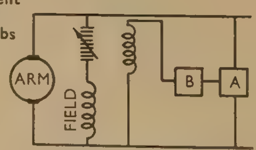

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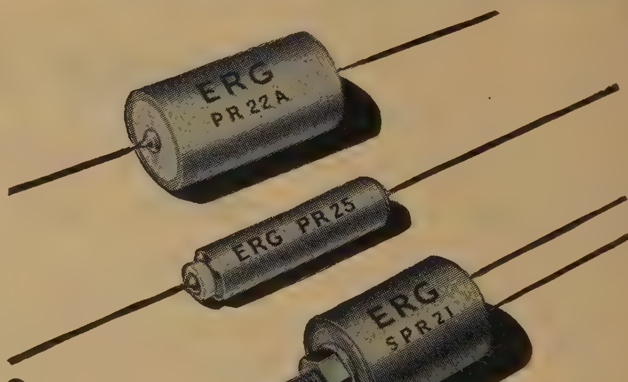

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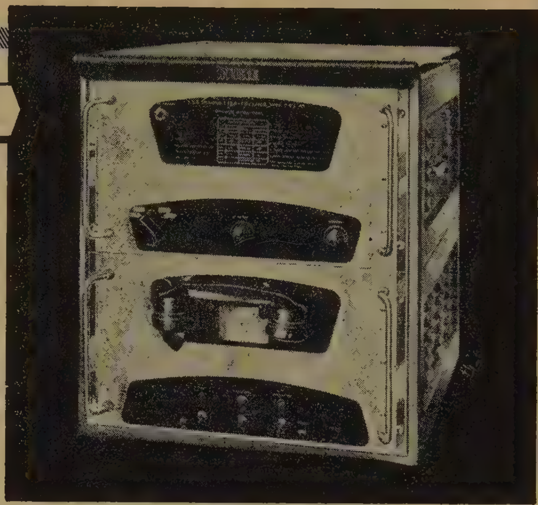

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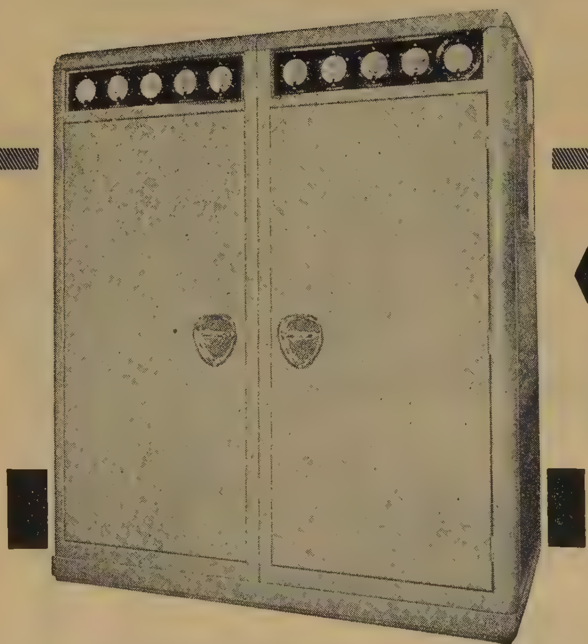

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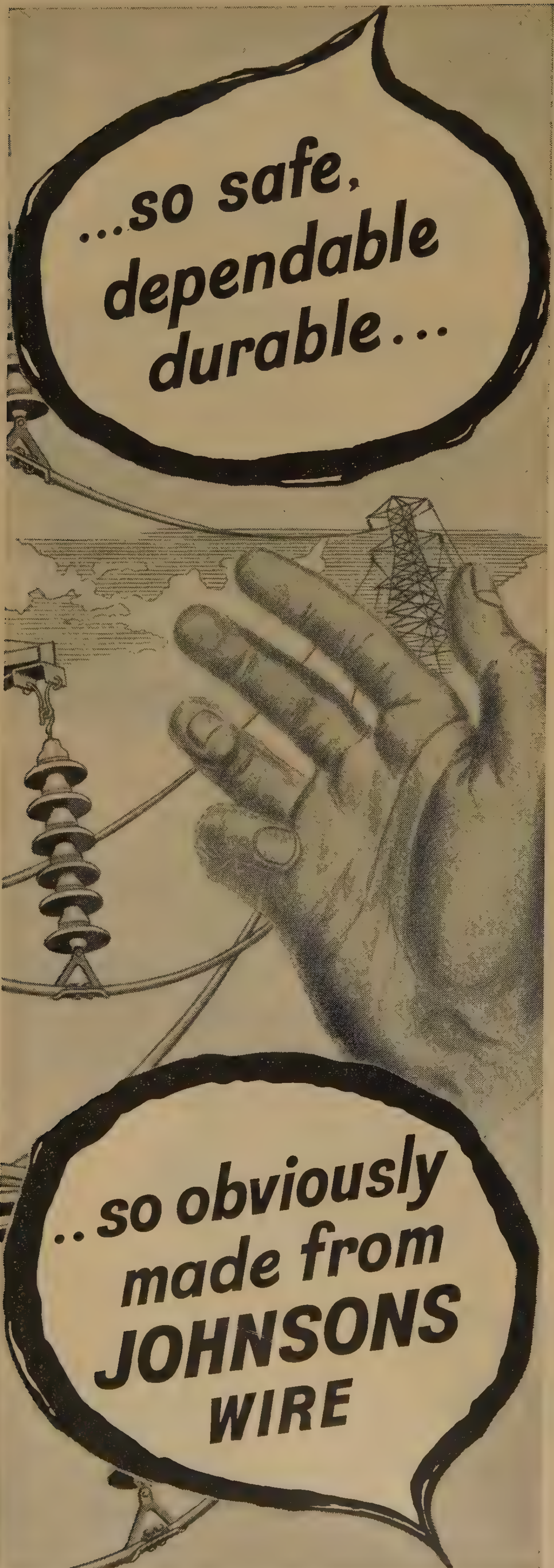

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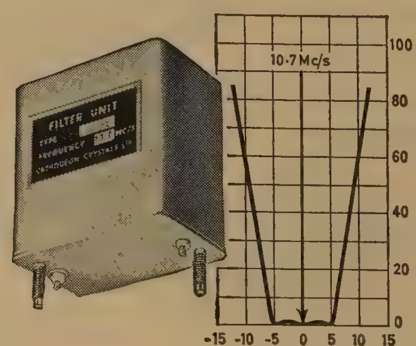

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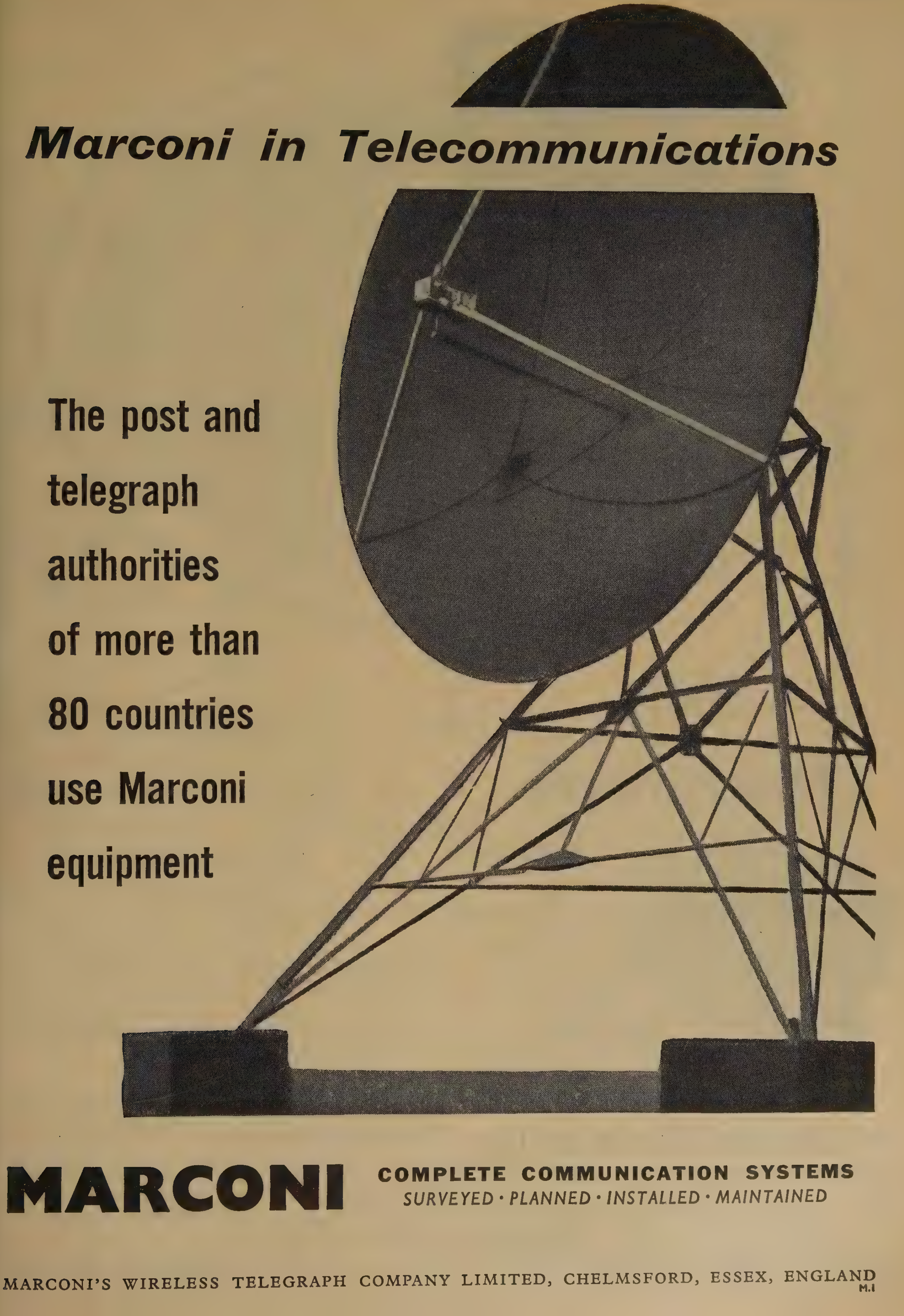
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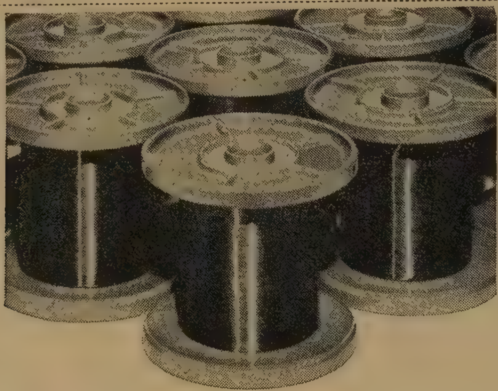

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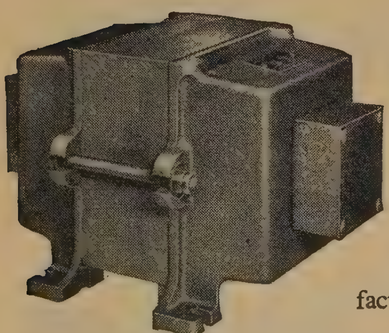

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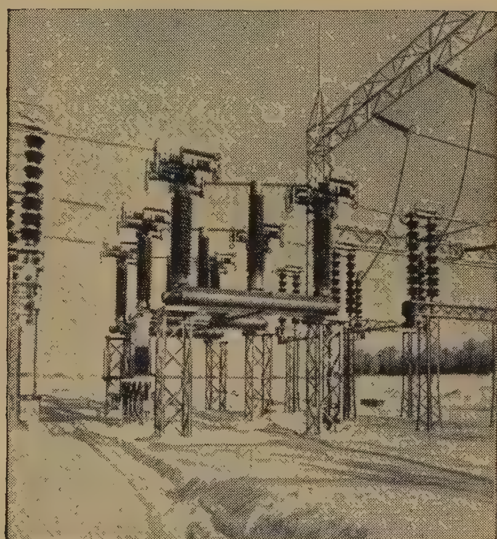

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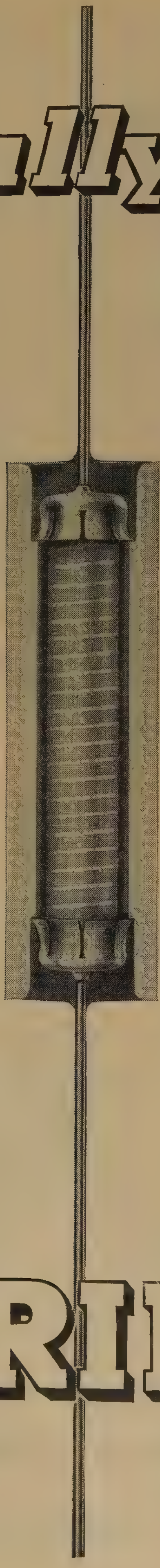

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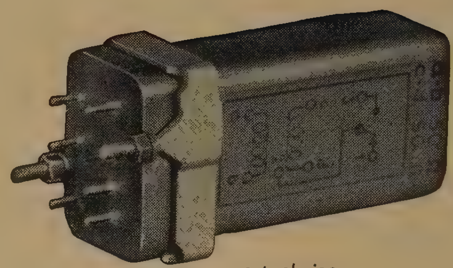

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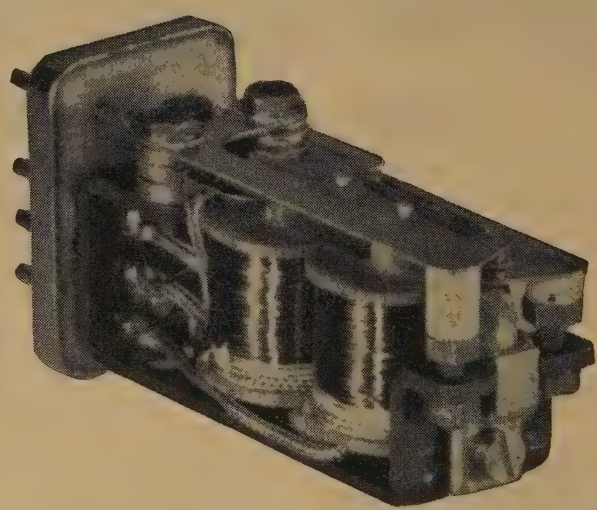

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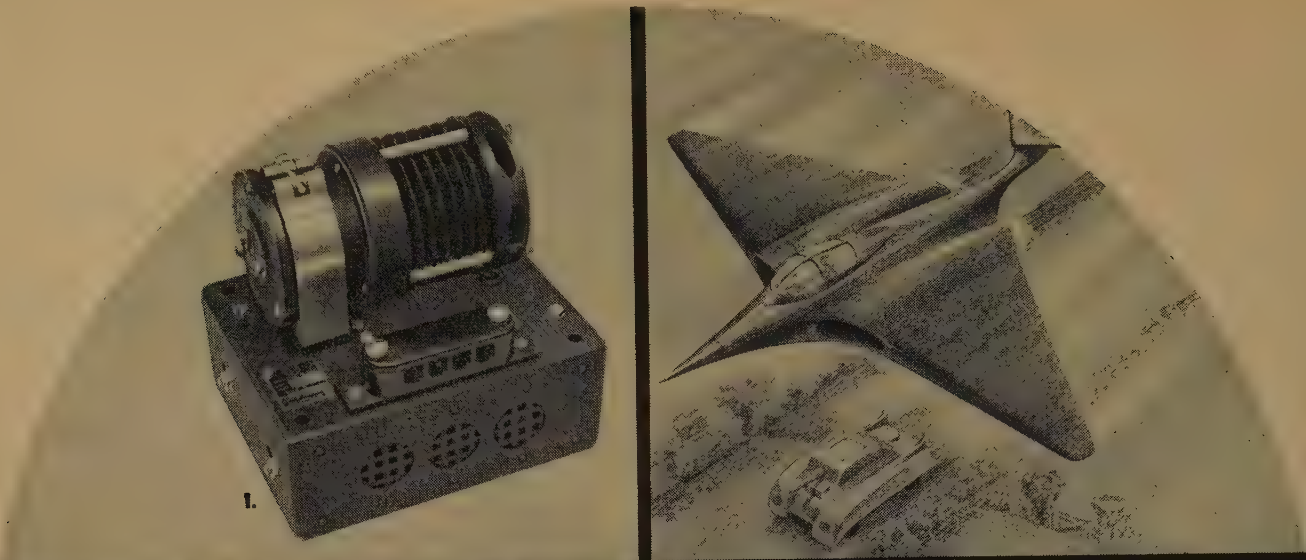

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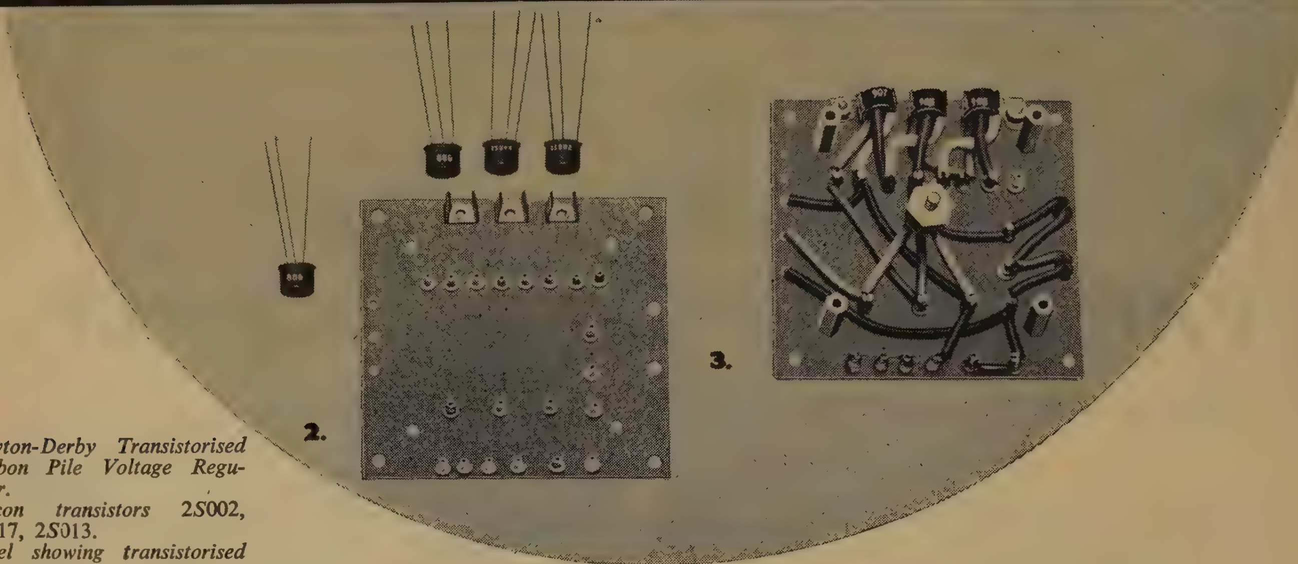

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THE SPECTRAL DENSITY OF THE A.M. NOISE IN REFLEX KLYSTRONS

By H. HÄGGBLOM, Tekn.lic.

(*The paper was first received* 29*th January, and in revised form* 25*th April*, 1959.)

SUMMARY

The a.m. noise spectrum in reflex klystrons is treated experimentally and theoretically. The frequency range of the measurements is 1–100 kc/s from the signal frequencies, which are 4·7 and 9·3 Gc/s. It is shown that the measurement equipment cannot consist of a crystal detector immediately followed by a low-frequency amplifier because the crystal noise up to about 20 kc/s is greater than the klystron noise. To reduce the effect of crystal noise the klystron is frequency modulated and the measurement equipment consists of a crystal diode as first detector, an i.f. amplifier, a vacuum diode as second detector and a selective l.f. amplifier. The measurements show that the noise density is, with great accuracy, constant in the actual range. This agrees with the theory, which is based on a treatment by Knipp. The deviation from Knipp's noise formula is due to the introduction of an inherent synchronizing effect in the noise current. The theory indicates that the noise is approximately constant up to a displacement of 10^{-4} to 10^{-3} from the signal frequency.

LIST OF SYMBOLS

C = Equivalent capacitance of the resonator.
e = Charge of the electron.
f = Oscillation frequency of the klystron.
f_0 = Resonant frequency of the cavity.
f' = Noise frequency under consideration.
$f'' = f' - nf$, where n is an integer.
$f''' = 2f_0 - f'$.
f_1 = Frequency of oscillator 1 (Fig. 1).
f_c = Frequency of oscillator 3 (Fig. 1).
$f_N = |f - f'|$.
G_e = Electronic conductance of the klystron.
$I(t)$ = Electron current as a function of time.
I_0 = D.C. component of $I(t)$.
$I^+(0, \tau')$ = Component of I passing the resonator gap in the direction towards the reflector (positive direction) at time τ'.
$I^+_{\omega'}$ = Spectral component of $I^+(0, \tau')$ at the angular frequency ω'.
$\overline{|I_1|^2}$ = Mean square of the noise current density in the resonator gap.
J_0, J_1, J_n = Bessel functions of the 0th, 1st and nth orders.
k = Relation between the cathode current and the reflected current.
m_f = Frequency-modulation index of the klystron caused by oscillator 1 (Fig. 1).
m_c = Frequency-modulation index of the klystron caused by oscillator 3 (Fig. 1).
n = Summation index.
$P_{Nf'}$ = Spectral density of the noise power.
Q = ω_0 times the stored energy in the cavity divided by the power of the loaded klystron.
T = Constant part of the electron transit time from the gap to the reflector space and back to the gap.
t = Time co-ordinate.
V = Resonator-gap voltage at the oscillation frequency.
V_0 = Accelerator voltage.
$V_{\omega'}$ = Spectral density of the voltage at angular frequency ω'.
$V^*_{\omega'}$ = Conjugate value of $V_{\omega'}$.
$v(\tau')$ = Electron velocity at time τ'.
v_0 = Constant part of the electron velocity at a given position.
$v_{\omega'}$ = ω'-component of the electron velocity at a given position.
X = 'Bunching parameter' of the klystron.
$Y_{\omega'}$ = Resonator admittance at angular frequency ω'.
τ' = Time at which an electron has reached the gap in positive direction of flight.
θ' = Transit angle at frequency f'.
$\omega, \omega_0, \omega'$, etc. = 2π times the frequency corresponding to the index.

(1) INTRODUCTION

The method of measuring a.m. noise in microwave oscillators is well known and reliable when the noise band is at least several megacycles per second from the signal frequency. The measured values are then in accordance with the theoretical ones obtained by Knipp.[4] Measurements in the neighbourhood of the signal

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Mr. Häggblom is with Aktiebolaget Svenska Elektronrör, Sweden.

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frequency must be done by observing the fact that crystal diodes have a very high noise spectral density at low frequencies. In connection with the synchronization of oscillators, Aitchison[1] described a method for noise measurements in which the first mixer was a coaxial diode. As he used an i.f. stage and a second detector, a crystal mixer in the first stage was also possible. On the other hand, Mueller[2] earlier published an experimental determination of a.m. noise in reflex klystrons using a crystal detector and in which no mention was made of how the influence from the crystal noise is avoided. In spite of this, his results agree, at least qualitatively, with those of Aitchison. This indicates that Mueller has also considered the influence of the crystal diode.

The experimental curve for the noise spectral density in the neighbourhood of the signal frequency does not fit Knipp's formula. Knipp derived, for the noise power density, the expression

$$P_{Nf'} \simeq \frac{F(f')}{(f'-f)^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where $F(f')$ is a slowly varying function of f'. Obviously, the expression is not valid for $f'=f$. Furthermore, it is not consistent with measurements at low values of $(f'-f)$. Indeed, Aitchison has measured constant power density when $(f'-f)$ was varied over the range 1–80 kc/s. It is thus important to find a correction to eqn. (1) for small values of the denominator. A method of obtaining such a correction is shown in Section 3. In Section 2 a method for noise measurement is described which is somewhat different from that used in Aitchison's paper.

(2) EXPERIMENTAL DETERMINATION OF NOISE IN REFLEX KLYSTRONS

The noise power density may be measured with the aid of a balanced mixer and two r.f. oscillators if they are identical or if one of them has negligible noise. The method is described by Aitchison. However, a method using a single oscillator is also possible. The noise spectrum of two types of reflex klystron was measured using two such methods. The simplest equipment consists of the oscillator under test, a silicon diode and a selective amplifier which is tuned to the frequency difference between the signal and the noise band. The second equipment is shown in Fig. 1.

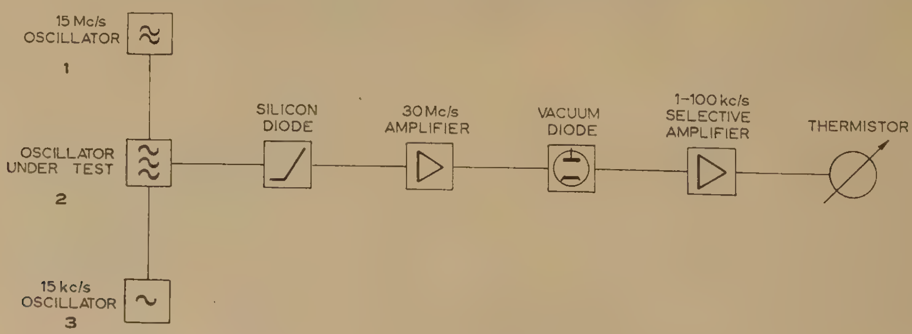

Fig. 1.—Equipment for noise measurement with two mixers.

The oscillator 1 is frequency modulating the reflex klystron 2. We assume for simplicity that the output from the klystron without frequency modulation can be expressed as

$$V_{out} = V\left(1 + \frac{2V_{\omega'}d\omega' \sin \omega_N t}{V}\right) \sin \omega t$$

The oscillation amplitude is here modulated with the angula frequency ω_N. The result is two a.m. noise bands at th angular frequencies ω' and $2\omega - \omega'$.

The frequency f_1 affects only ω. Some of the resulting spectr components are as follows

$$V\left(1 + \frac{2V_{\omega'}d\omega'}{V} \sin \omega_N t\right)\Big\{J_0(m_f) \sin \omega t$$
$$+ J_1(m_f)[\sin(\omega + 2\pi f_1)t - \sin(\omega - 2\pi f_1)t]$$
$$+ J_2(m_f)[\sin(\omega + 4\pi f_1)t + \sin(\omega - 4\pi f_1)t] + \ldots\Big\}$$

The crystal diode can be considered as a squaring device, an among the squared components we amplify those of the form

$$V^2\left(1 + \frac{4V_{\omega'}d\omega'}{V} \sin \omega_N t\right)[J_0(m_f)J_2(m_f) + J_1^2(m_f)] \cos 4\pi f_1 t$$

The square of the a.m. noise is omitted. The i.f. amplifier tuned to $2f_1$ or 30 Mc/s. The voltage is amplitude modulate and the modulation index is proportional to $V_{\omega'}$, i.e. the spectr density at the angular frequency ω'. The i.f. amplifier is followe by a vacuum diode and an amplifier which is tuned to $|\omega' - \omega$

The oscillator 3 is also applied for frequency modulation the signal. Its purpose is to simulate noise so that we ca calibrate the equipment. It is not important whether amplitud or frequency modulation is employed, but frequency modulatio is chosen for convenience. We write, for the calibrator, th angular frequency ω_c and modulation index m_c. Omitting th noise, the input to the crystal contains components of the for

$$J_0(m_c)J_2(m_f)\ [\sin(\omega + 2\omega_1)t + \sin(\omega - 2\omega_1)t]$$
$$+ J_0(m_f)J_2(m_c)\ [\sin(\omega + 2\omega_c)t + \sin(\omega - 2\omega_c)$$

After squaring, we remove the components

$$J_0(m_f)J_0(m_c)J_2(m_f)J_2(m_c)\ [\cos 2(\omega_1 + \omega_c)t + \cos 2(\omega_1 - \omega_c)t$$

Together with the i.f. carrier we get an a.m. signal, which ca be detected and used by calibration of the equipment.

The noise measurements are done at 4·7 and 9·3 Gc/s. Th former frequency is generated by a klystron having a maximu power output of 150 mW. At the higher frequency the klystr type VA201 is used. Two methods are used to investigate th error caused by the crystal noise when using the simpl measuring equipment. The result is shown in Fig. 2. Th mean oscillation frequency is f_0. The signal/noise ratio is $\frac{1}{2}$ because it is referred to the two symmetrical noise band

Curve (i) shows the measured values at 4·7 Gc/s when the amplifier immediately follows the silicon diode. Curve (shows the same quantity at 9·3 Gc/s measured with the equi ment shown in Fig. 1. Both Mueller and Aitchison ha found constant noise density between 1 and 10 kc/s. By mea of reflex klystrons Dalman and Rhoads[3] obtained a signal/no

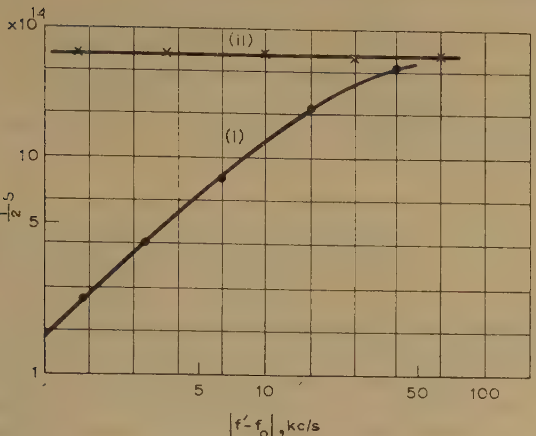

Fig. 2.—Measured signal/noise ratio as a function of frequency.

(a) The measurement equipment contains no i.f. amplifier.
(b) The measurement equipment contains an i.f. amplifier for 30 Mc/s.

ratio of the order of 10^{15} at 10 Mc/s from the signal frequency. As the quantity is also of the same order at 10 kc/s, it is reasonable to assume that it is approximately constant over the whole range 1 kc/s–10 Mc/s. This result is also consistent with the following theory. We then conclude that curve (i) in Fig. 1 does not explain the signal/noise ratio of the klystron. Its inclination is apparently due to crystal noise. In fact, it is well known that the crystal noise density is inversely proportional to the frequency over the actual range. Thus the crystal noise is greater than the klystron noise up to about 20 kc/s.

A second result of the measurements is that the a.m. noise in the neighbourhood of the signal frequency seems to be at least approximately independent of the signal frequency. This is apparent from curves (i) and (ii) at $(f' - f_0) \simeq 50$ kc/s, where the crystal noise is less important. Here the signal/noise ratio at 4·7 and 9·3 Gc/s are of the same order, which also agrees with the theory.

(3) THEORETICAL TREATMENT OF A.M. NOISE IN REFLEX KLYSTRONS

The calculations are based on Knipp's theory. For clarity we write down his mathematical assumptions and two connected equations which follow. By solving the equations, we must consider a kind of synchronization between the signal frequency and the h.f. components in the electron current. This is neglected by Knipp.

To make the expressions as short as possible we neglect the transit time in the resonator gap and consider only the transit time in the reflector space. When all r.f. fields are absent this is termed T. The theory can, without difficulty, be extended to the case when the transit time in the gap is greater than zero.

It is assumed that an electron passes the gap at time T'. We consider the electron current and velocity in the gap which at this moment are directed toward the reflector. For these quantities we have

$$\left.\begin{aligned} I^+(0,\tau') &= -I_0 + \int_{-\infty}^{\infty}\frac{d\omega'}{2\pi} I^+_{\omega'}\varepsilon^{j\omega'\tau'} \\ v(\tau') &= v_0 + \int_{-\infty}^{\infty}\frac{d\omega'}{2\pi} v_{\omega'}\varepsilon^{j\omega'\tau'} \end{aligned}\right\} \quad . \quad . \quad (1)$$

Across the gap we have the voltage

$$V(t) = V\sin\omega t + \int_{-\infty}^{\infty}\frac{d\omega'}{2\pi}V_{\omega'}\varepsilon^{j\omega' t} \quad . \quad . \quad . \quad (2)$$

The transit time of the electrons at arbitrary positions and when the r.f. field is present can be calculated from eqns. (1) and (2) with the aid of Lorentz's force equation. From the transit time Knipp has calculated the total electron current in the gap. The current is then eliminated using the admittance equation:

$$I(t) = -(1-k)I_0 + \sum_{n\neq 0} Y_{n\omega}V_{n\omega}\varepsilon^{jn\omega t} + \int_{-\infty}^{\infty}\frac{d\omega'}{2\pi}Y_{\omega'}V_{\omega'}\varepsilon^{j\omega' t} \quad (3)$$

kI_0 is that part of the direct current which is reflected through the gap. The second term on the right-hand side represents the possible signal frequencies and the last term represents the noise current. We assume that the tube oscillates at the frequency f_0 at which the resonator has zero susceptance. For the noise frequency f' we have

$$Y_{\omega'} = -G_e + 2j(\omega' - \omega_0)C \quad . \quad . \quad . \quad . \quad (4)$$

$$\left.\begin{aligned} G_e &= -\frac{2kI_0 J_1(X)}{V} \\ X &= \frac{\omega_0 TV}{2V_0} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad (5)$$

Knipp shows that the noise voltage component, $V_{\omega'}$, is a function of the noise current at ω' and at $\omega'' = \omega' - n\omega_0$ and of the component $V_{\omega'''}$ at $\omega''' = 2\omega_0 - \omega$. ω''' is the mirror angular frequency of ω'. We then arrive at the following equations:

$$\left.\begin{aligned} \left(\frac{jA}{2} - B\right)V_{\omega'} - j\frac{A}{2}V^*_{\omega'''} &= j(I^+_{\omega'} - kI^-_{\omega'})^* \\ j\frac{A}{2}V_{\omega'} + \left(B - \frac{jA}{2}\right)V^*_{\omega'''} &= -j(I^+_{\omega'''} - kI^-_{\omega'''}) \end{aligned}\right\} \quad . \quad (6)$$

where

$$A = \frac{2kI_0XJ_2(X)}{V}$$

$$B = 2(\omega' - \omega_0)C$$

$$I^-_{\omega'} = \sum_n J_n(X)\varepsilon^{-jn\pi}\left(I^+_{\omega''} + jI_0\theta'\frac{v_{\omega''}}{v_0}\right)_{\omega''=\omega'-n\omega_0}$$

$$\theta' = \omega' T$$

Knipp has solved eqn. (6) for $V_{\omega'}$. This quantity is then a function of $I^+_{\omega'}$, $I^+_{\omega''}$ and $I^+_{\omega'''}$. Taking the mean square of $V_{\omega'}$ all terms containing phases of currents vanish. However, this operation leads to an inconsistency. An alternative method of solving eqn. (6) is to multiply both equations by their conjugates, respectively, and to take the mean of these quantities. Those terms containing phases of $V_{\omega'}$ and $V_{\omega'''}$ disappear. If we then solve for $|V_{\omega'}|^2$ the result will not be the same as that of Knipp.

The explanation of the different result is apparently hidden in the influence of the phases of the r.f. quantities on the physical behaviour. Let us consider an electron beam which is modulated only by the angular frequency ω'. This r.f. current causes a corresponding gap voltage, and if $|\omega' - \omega_0|$ is sufficiently small the oscillator must be synchronized to ω'. This is also clear from eqn. (6) if we solve the equations for the phase of $V_{\omega'}$. For a given value of $|V_{\omega'}|$ the phase will be imaginary when ω' approaches ω_0. This behaviour characterizes an instability.

In a strictly mathematical sense, the correct way to solve the problem is to take account of the synchronization. But this is a transient phenomena, and, furthermore, the synchronizing source is changing statistically. Thus the problem is very involved. However, there is one possibility of avoiding synchronization at ω' in the stationary state, namely if $|V_{\omega'''}| = |V_{\omega'}|$ and their phase difference is appropriate. $V_{\omega'''}$ will then balance the synchronizing effect of $V_{\omega'}$. The incoming noise current in the klystron is very slowly varying with the frequency. This is equal to the statement that $|V_{\omega'''}| = |V_{\omega'}|$.

The phases are determined by eqns. (6) and do not have values such that synchronization is balanced out. In this respect the system is non-stationary and the physical effect is therefore only that there is f.m. noise in the klystron. However, neglecting f.m. noise one can consider a system with the appropriate phases. This is the same thing as taking the mean square of eqn. (6) directly. Thus we get

$$\frac{\overline{|V_{\omega'}|^2}}{V^2} = \frac{\overline{|I_1|^2}/I_0^2}{2k^2X^2[\mathrm{J}_2(X)]^2 + \dfrac{4V^2(\omega' - \omega_0)^2C^2}{I_0^2}} \quad . \quad (7)$$

where $I_1 = I^+_{\omega'} - kI^-_{\omega'}$.

We now introduce the Q-factor of the loaded resonator. For the spectral density of the noise power we write the symbol $P_{Nf'}$.

Then

$$P_{Nf'} = \frac{2V_0\overline{|I_1|^2}XJ_1(X)}{k\theta I_0\left\{X^2[\mathrm{J}_2(X)]^2 + \dfrac{8Q^2[\mathrm{J}_1(X)]^2(\Delta\omega)^2}{\omega_0^2}\right\}} \quad . \quad (8)$$

where $\Delta\omega = \omega' - \omega_0$

When $\Delta\omega$ approaches zero we have

$$P_{Nf_0} = \frac{2V_0\overline{|I_1|^2}}{k\theta I_0 X[\mathrm{J}_2(X)]^2} \quad . \quad . \quad . \quad . \quad . \quad (9)$$

We can easily compute the value of $\Delta\omega$ when $P_{Nf'}$ has fallen to half its maximum value. If $Q\sqrt{2} = 1\,000$ and $X = 1$ we have $\Delta\omega/\omega_0 = 1 \cdot 3 \times 10^{-4}$. For greater bunching the parameter $\Delta\omega$ increases.

A simple way to compute the noise density from eqn. (8) or (9) is to consider $\overline{|I_1|^2}$ as full shot noise:

$$\overline{|I_1|^2} = 2eI_0$$

This is done using the following assumptions:

$$f_0 = 10\,\mathrm{Gc/s}$$
$$V_0 = 250\ \mathrm{volts}$$
$$k = 0 \cdot 9$$
$$\theta = 27/4$$
$$X = 2$$

These values are typical for the reflex klystron type VA201B, for which the signal/noise ratio of 155 dB was measured at small values of $\Delta\omega$. The klystron thus gives the signal power of 30 mW. Computation of P_{Nf_0} from eqn. (9) gives the noise density $5 \cdot 6 \times 10^{-17}$ watt per c/s. The corresponding signal/noise ratio is 152 dB.

(4) CONCLUSIONS

The a.m. noise in reflex klystrons is mainly a function of the difference between the signal frequency and the noise band under consideration. Measurements using independent methods have been carried out by Mueller, Aitchison and the author. All these show that the noise density is nearly independent of Δf, at least within a relative bandwidth of 10^{-5}. Since the noise in a crystal diode is inversely proportional to frequency down to about 1 kc/s, both noise sources can be easily separated. Furthermore it is shown that the noise power is approximately independent of the signal frequency when Δf is small.

The theoretical treatment is based on the theory by Knipp according to which the noise power is inversely proportional to Δf^2. This formula is not in accordance with experience for small values of Δf. A more consistent formula is derived by considering the synchronizing effect of the r.f. components in the electron beam. The theory then gives a nearly constant noise spectrum within a relative bandwidth of about 10^{-4}–10^{-3}. Furthermore, it is suggested that the synchronizing effect is also a source of f.m. noise.

(5) ACKNOWLEDGMENT

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HIGH-QUALITY MICROPHONES

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SUMMARY

The paper reviews present-day practice in the design and operation of high-grade microphones used for broadcasting, sound reinforcement, sound recording, acoustical measurements, etc.

General requirements, basic operating principles and design details are given for typical microphones of various types. They are classified acoustically into omnidirectional and directional categories for the purposes of discussion and analysis.

LIST OF SYMBOLS

p = Sound pressure.
K = Fraction of the total output of a microphone due to the pressure-gradient component.
θ = Angle of sound-wave incidence relative to a microphone axis.
D = Diameter.
d = Effective distance corresponding to the sound pressure-gradient force experienced by a microphone.
λ = Wavelength of sound.
v = Velocity of propagation of a sound wave.
x = Displacement.
$\dot{x}$ = Linear velocity.
A = Area.
M = Mass.
S = Stiffness.
R = Resistance.
Z = Impedance.
C = Compliance (=1/S).
(M, S, R, Z, C: Expressed in analogous acoustic units.)
B = Flux density of a magnetic field, gauss.
ϕ = Phase shift between two sound pressures.

(1) INTRODUCTION

(1.1) Requirements to be Met by Microphones

Apart from particular requirements imposed by special applications, it is possible to list some desirable properties which microphones should possess. The following general requirements for high-grade microphones are suggested:

(*a*) The frequency response should be as flat as possible over an audio-frequency range of 35 c/s to 15 kc/s, approximately, and the shape of the response curves should be as nearly identical as possible for different samples of the same product. This is particularly important in stereophonic applications and where any electrical equalization is to be carried out. Tolerances of about $\pm 1\cdot 5$ dB are suggested for flatness and matching. Wider tolerances may be permitted towards the upper and lower extremes of the frequency range.

(*b*) The polar response should be either omnidirectional or suitably directional, and in either case the shape of the curve should have the smallest possible variation with frequency.

(*c*) The transient response should be satisfactory, and in particular should be free from any 'ringing' or prolongation of the decay at any frequency. This fault may occur if, owing to a 'break-up' resonant mode of vibration, some part of the moving system oscillates in a lightly damped condition. The contribution of the element in question may not be noticeable in the steady-state frequency response, but it may introduce transient distortion.

(*d*) The electro-acoustic conversion efficiency of microphones should be sufficiently high to ensure a good signal/noise ratio. In general, the output should not be less than -88 dB/volt per dyne/cm^2 at 30 ohms.

Non-linear distortion should be low enough to be comparable with that of a well-designed amplifier chain and should be less than $0\cdot 5\%$ total harmonics at the highest sound levels, e.g. $+110$ to $+120$ dB relative to the threshold. Output impedances should be suitable and should correspond to established practice (e.g. 30, 300 ohms, etc.).

(*e*) Microphones should not be unduly affected by wind, breath, mechanical shocks and vibration, electric or magnetic fields, or by high-intensity lighting and its associated radiant heat.

(*f*) Unfavourable climatic conditions may have to be withstood, and the entry or collection of dust particles should be resisted.

(*g*) The size and weight of microphones should be moderate and the appearance unobtrusive.

(*h*) The mountings, electrical connectors, cables, etc., should be robust, convenient and non-microphonic in normal use.

There must be considerable compromise in a given design with regard to many of these requirements. For instance, it is not possible at present to obtain an entirely accurate polar-response shape at high frequencies unless the unit is made so small that a serious loss of electro-acoustic conversion efficiency is encountered. Also, different applications may result in a shift of the emphasis which is placed on particular requirements.

(1.2) General Properties of Omnidirectional and Directional Microphones

Microphones can be broadly classified acoustically as those with omnidirectional polar response and those with directional response. Omnidirectional types are used where sound has to be collected over a large solid angle without appreciable directional discrimination. For example, they are used where the sound from widely dispersed sources is to be picked up, where the maximum ratio of reverberant to direct sound is desired, and in the measurement of total noise or of reverberant sound fields. Omnidirectional microphones are basically pressure-operated devices, since they are responsive to the acoustic sound pressure, which is a scalar, i.e. a non-directed, quantity.

Thus an omnidirectional microphone has a construction characterized by a rigid, non-porous, enclosed case with the diaphragm or other movable part of the transducer occupying part of the surface. The overall size must be small compared with the wavelength of sound; otherwise the effects of sound-

This is an 'integrating' paper. Members are invited to submit papers in this category, giving the full perspective of the developments leading to the present practice in a particular part of one of the branches of electrical science.
Mr. Gayford is with Standard Telephones and Cables, Ltd.

Table 1A

WAVE DIFFRACTION LOSS ON CYLINDER END RELATIVE TO NORMAL WAVE INCIDENCE (0°)

Frequency	Diameter $D = 0{\cdot}5$ in			Diameter $D = 1{\cdot}0$ in		
	D/λ	Max. loss	Loss at 90°	D/λ	Max. loss	Loss at 90°
kc/s		dB	dB		dB	dB
1·0	0·037	0·5	0·5	0·074	1·5	1·5
5·0	0·185	3·0	2·5	0·37	6·5	5·0
10·0	0·37	6·5	5·0	0·74	13·0	8·5
12·0	0·45	8·5	6·5	0·90	14·0	9·5
15·0	0·55	10·0	7·5	1·1	15·5	9·5

Table 1B

WAVE DIFFRACTION LOSS ON SPHERE. (RELATIVE TO 0°)

Frequency	Diameter $D = 0{\cdot}5$ in			Diameter $D = 1{\cdot}0$ in		
	D/λ	Max. loss	Loss at 90°	D/λ	Max. loss	Loss at 90°
kc/s		dB	dB		dB	dB
1·0	0·037	0	0	0·074	0	0
5·0	0·185	1·2	1·0	0·37	5·0	3·5
10·0	0·37	5·0	3·5	0·74	7·2	4·5
12·0	0·45	5·7	4·0	0·90	9·0	4·7
15·0	0·55	6·2	4·2	1·1	10·0	5·0

wave diffraction round the case will cause the sound pressure on the diaphragm surface to change with the angle of incidence.

Tables 1A and 1B illustrate the relationship which holds between size, shape and diffraction effects. It will be seen that, even for the smallest generally practicable sizes (about 0·5 in diameter), there are still diffraction effects of up to 10 dB. An additional drop, the phase-loss effect, occurs at high frequencies for sound incident in a direction other than normal to the plane of a diaphragm, because there is an appreciable phase difference in the pressure wave over the distance represented by the diaphragm diameter. The magnitude of phase-loss effects for typical diaphragm sizes is given in Table 1C. It is therefore

Table 1C

PHASE LOSS EFFECTS ACROSS CIRCULAR DIAPHRAGMS FOR SOUND INCIDENT AT 90°. (RELATIVE TO 0°)

Frequency	Diameter $D = 0{\cdot}5$ in		Diameter $D = 1{\cdot}0$ in		Diameter $D = 1{\cdot}5$ in	
	D/λ	Phase loss	D/λ	Phase loss	D/λ	Phase loss
kc/s		dB		dB		dB
5·0	0·18	0·3	0·37	0·7	0·55	1·5
10·0	0·37	0·7	0·74	2·7	1·11	6·5
12·0	0·45	1·0	0·89	4·0	1·33	9·0
15·0	0·55	1·5	1·11	6·5	1·67	11·0

necessary, even for small microphones, to resort to special means in order to achieve a good omnidirectional response.

Microphones with directional properties depend fundamentally for their operation on effects dependent on distance. The pressure-gradient-operated types are responsive to the pressure difference between two separate areas. This is a vector quantity aligned with the direction of sound-wave propagation. Other types depend on reflectors or on the effects of wave propagation in various structures in order to achieve directivity.[1]

Directional microphones are very valuable in broadcasting and sound recording when it is desired to discriminate against excessive reverberant sound or background noise; they enable a microphone to be used at a greater distance from a sound source; and may give a sense of directional sound location, as in some stereophonic sound-recording systems.[2]

In addition to obtaining the desired polar response, the relationship between the operating principles of the mechano-electric transducer and the motion imparted to the moving element must be studied, so that a flat frequency response results. To a first order of accuracy, the relationships which apply to the various basic types of microphone transducer are shown in Fig. 1.

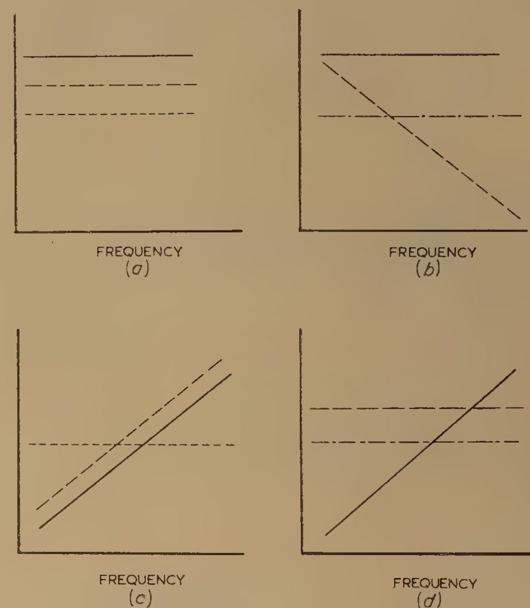


Fig. 1.—Basic acoustic-mechanical operating principles of microphones.

The ordinate scales represent the following quantities:
——— Actuating force. — — — — Mechanical impedance.
- - - - Velocity. —·—·— Displacement.

(*a*) Pressure-operated system with resistance control, e.g. omnidirectional moving-coil microphone.
(*b*) Pressure-operated system with stiffness control, e.g. omnidirectional piezo-electric or condenser microphone.
(*c*) Pressure-gradient-operated system with mass control, e.g. bidirectional or uni-directional ribbon microphone or unidirectional moving-coil microphone.
(*d*) Pressure-gradient-operated system with resistance control, e.g. unidirectiona condenser microphone.

For (*a*) and (*c*), constant velocity of the conductor in the polarizing magnetic fiel generates constant e.m.f. at all frequencies. For (*b*) and (*d*) constant displacement o the system generates constant e.m.f. at all frequencies.

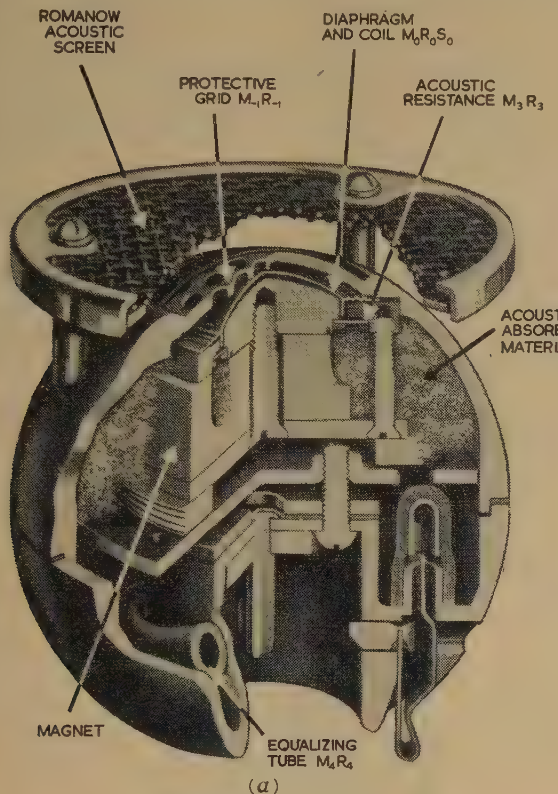
Thus, in each case, the required mechanical-impedance charac-teristic of the system is derived. In Fig. 1(*a*) the governin element in an omnidirectional moving-coil microphone is a acoustic resistance tightly coupled to the diaphragm system which should have a relatively low restoring stiffness. In Fig 1(*b*) a stiffness-controlled system is produced by the light an relatively tightly stretched diaphragm of the condenser micro phone or by the high stiffness of a crystal. For a simple first order gradient type, such as the ribbon microphone, the actuatin force is approximately proportional to the frequency, and th mechanical impedance must be mass controlled. The ribbon i therefore made very flexible and has its fundamental resonanc at the lowest audio frequencies [see Fig. 1(*c*)]. Fig. 1(*d*) show that a gradient-operated condenser microphone must be resistance-controlled system.

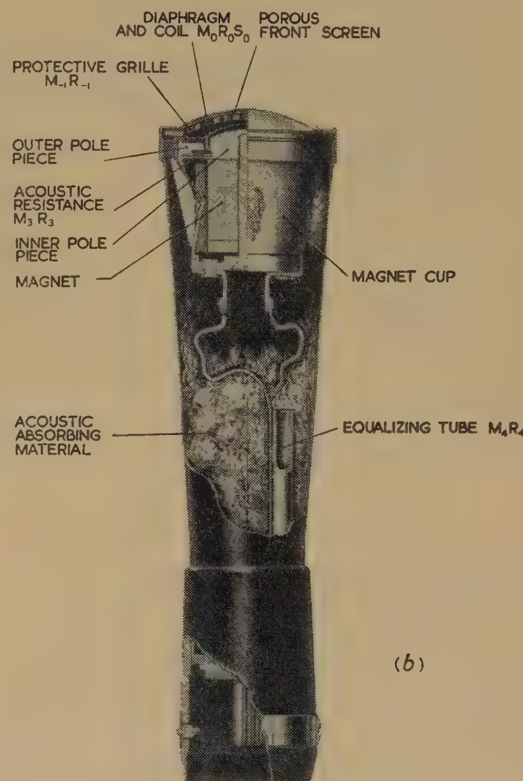
The design of practical microphones of high performance is necessarily more complicated than is indicated by this simplified approach. In the following Sections, examples representative of microphones at present in use will be described in more detail.

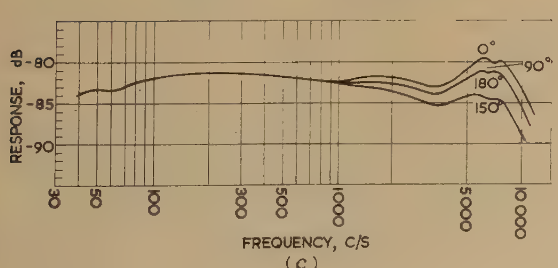
(2) OMNIDIRECTIONAL MICROPHONES

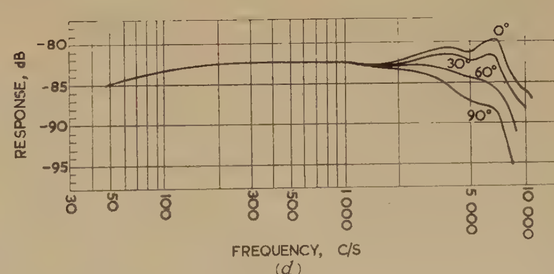
(2.1) Conventional Free-Field Omnidirectional Microphones

Figs. 2(*a*)–(*e*) show typical omnidirectional moving-coil microphones and their response curves. The diaphragms, which are made of aluminium alloy or a suitable plastic, consist of a stiff-domed centre carrying a self-supporting speech coil wound from aluminium tape. The surround has tangential corrugations which give higher compliance and less amplitude non-linearity than conventional annular corrugations. The effective area and shape of the diaphragm are a compromise between size and freedom from 'break-up' resonances, together with the need for adequate sensitivity. The front protective grille, the pole-piece and magnet system providing a radial magnetic field across the speech coil, the finite volume of the









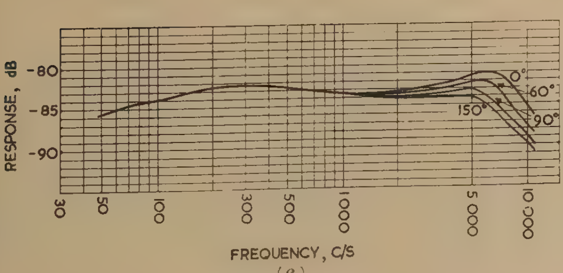


Fig. 2.—Typical omnidirectional moving-coil microphones and their response curves.

(*a*) An omnidirectional moving-coil microphone.
(*b*) A tubular omnidirectional moving-coil microphone.
(*c*) Free-field response of the microphone of (*a*).
(*d*) Free-field response of the microphone of (*b*) without a Romanow front screen fitted.
(*e*) Free-field response of the microphone of (*b*) with a Romanow front screen fitted.
For (*c*), (*d*) and (*e*), 0 dB = 1 volt per dyne/cm².

enclosing case and a leak path from the inside of the case (in order to prevent atmospheric changes from altering the mean position of the diaphragm relative to the pole-pieces)—all represent acoustic masses and compliances which are coupled to the diaphragm. In general, they may be regarded as forming Helmholtz resonators, and their resonant-frequency modes and damping must be arranged so that the system will present the correct mechanical impedance at any frequency when referred to the diaphragm. In addition, some cavities, such as the main-case volume, will be large enough at high frequencies to give rise to standing-wave effects which must be either damped out by suitably distributed acoustic absorbents or taken into account.

In the microphone of Fig. 2(*a*) the main case is spherical, as it is well established that diffraction and obstacle effects are less for a sphere than for other shapes. In the microphone of Fig. 2(*b*) the main case has the form of a tapering cylinder with as small a diameter as possible. In both cases a Romanow front screen[3] helps to maintain an omnidirectional response. The front screen for the tubular microphone can be omitted at the expense of the upper-frequency omnidirectional properties [see Figs. 2(*d*) and (*e*)]. This is often done to make the microphone as unobtrusive as possible in appearance. The Romanow screen consists of a layer of fabric of such a size and acoustic resistance that sounds arriving in the front hemisphere are attenuated whilst those arriving in the rear hemisphere are enhanced by reflection from the screen, thus enabling diffraction effects to be largely neutralized.

An analogous electrical circuit for these microphones is given in Fig. 3. It is acknowledged that the derivation and study of

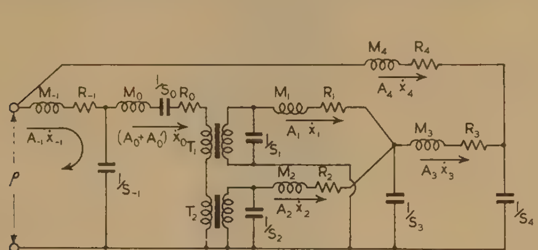


Fig. 3.—A lumped-constant analogous electrical circuit for the microphones of Figs. 2(*a*) and (*b*).

p = Acoustic pressure on microphone openings.
T_1 and T_2 are ideal transformers representing the action of the dome and surround areas of the diaphragm.
The other constants of the circuit will be readily identified from Figs. 2(*a*), 2(*b*) and the functional diagram of Fig. 14.

analogous electrical circuits leads to an easier understanding of the effect of the various elements and greatly aids development work. These circuits can be set up and measured much more readily than actual microphones (see Section 8.1). It is seen that the equations of motion for the acoustic-mechanical system also apply to the analogous electrical circuit. The various elements of the microphones are identified on the circuit of Fig. 3. The areas on the rear side of the diaphragm inside and outside the speech coil are coupled via the pole-gap slots on each side of the speech coil, and this configuration leads to the two transformers in the analogous circuit.

Wave-propagation effects are assumed to be absent in that lumped constants are used for the different elements. With the distributed damping used, the circuit represents a useful approximation to the actual microphone. We have seen that the characteristics of the acoustic-mechanical network for a pressure-operated moving-coil microphone have to be such that the diaphragm is driven at a constant velocity over the frequency range for constant acoustic incident pressure. The simple resistance control required can be simulated over a wide frequency range only by the adjustment of various elements to produce the desired response. For example, the leak tube is given a value such that it interacts with the case volume and other elements to maintain the response at extreme low frequencies. The front-grille openings form a damped Helmholtz resonator with the air cavity in front of the diaphragm and help to maintain the extreme top response. The governing element over the middle frequency range is the acoustic resistance below the coil slot.

Other omnidirectional microphones are operated on dynamic, electrostatic or piezo-electric principles.[4, 5, 6] The units generally have to be wider than about 0·5 in in order to have a useful efficiency under general free-field acoustic conditions, so that the response becomes directional at high frequencies. As alternatives to the Romanow front screen, three measures have been adopted to minimize this effect, namely (*a*) the use of near-spherical or smooth shapes to minimize diffraction effects, (*b*) the use of a centrally anchored diaphragm for condenser microphones, and (*c*) the use of front entries in the form of narrow annular peripherial slits around the front edge of the microphone.[7] The last two arrangements depend on the principle that diffraction effects are less severe around the edges of a front face than in the centre. Phase-loss effects for waves moving across the diaphragm may be made more severe but for fairly small units the gain due to reduced diffraction effects can outweigh any increase in phase-loss effects.

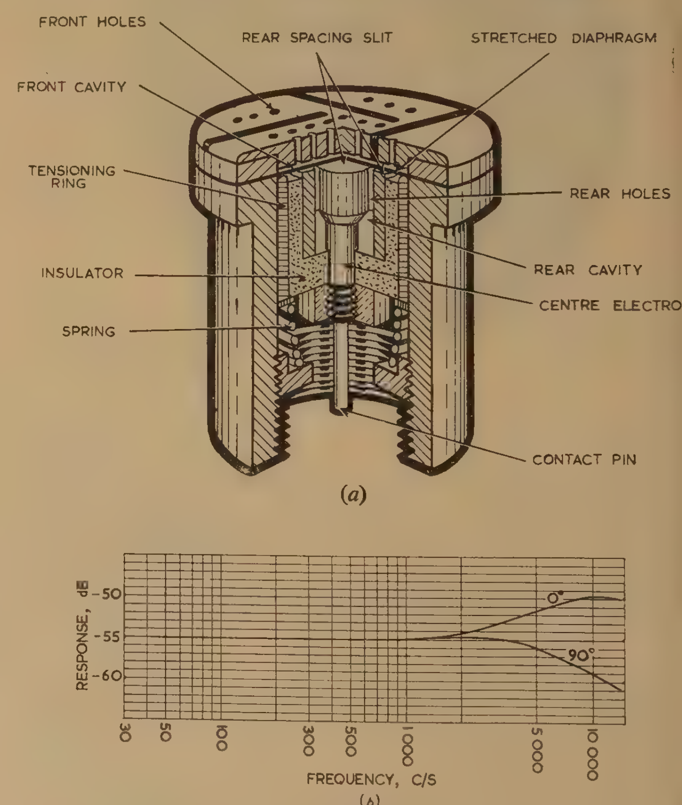


Fig. 4.—An omnidirectional condenser microphone.

(*a*) Cut-away view.
(*b*) Free-field response.
0 dB = 1 volt per dyne/cm².

Fig. 4(*a*) shows a small omnidirectional condenser microphone. The construction is made as robust as possible in order to give extremely stable characteristics. The stretched metal diaphragm is given a constant tension by spring loading on

peripheral ring. The central back electrode renders the diaphragm substantially aperiodic by means of acoustic resistance in the form of the spacing slit and the small holes communicating between the electrode face and a rear cavity. Fig. 4(*b*) shows the response curves.

(2.2) Probe-Tube Microphones

Probe-tube microphones comprise two categories of omnidirectional microphones. In the first category is the probe microphone designed to produce a small unobtrusive free-field sound-pressure pick-up point by using the end of a tube communicating with the main body of the microphone. One example is in the form of a floor or table stand, the moving-coil microphone unit being accommodated in the base with a tube of about $\frac{3}{8}$ in diameter rising to normal speaking height.[8] One of the main problems in the design of probe microphones is the elimination of marked standing-wave effects along the probe tube. In the above microphone, wave reflections are largely eliminated by terminating the tube with an acoustic resistance near that of the characteristic impedance of the air in the tube. To achieve this, an acoustic transformer is used at the end of the tube, since any practical acoustic resistance, such as a damped diaphragm, has a specific acoustic resistance exceeding that of air. The transformer consists of a flare or coupling unit connecting the diaphragm to the end of the tube, whilst the rear of the diaphragm is tightly coupled to a suitable acoustic resistance which terminates the tube via the acoustic transformer (see Fig. 5). A high-frequency boosting coupler unit terminates the open end.

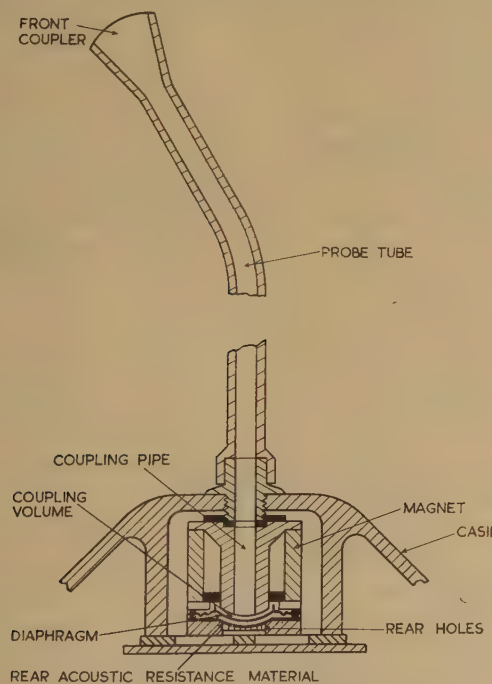


Fig. 5.—Free-field probe microphone.

The second category of probe unit includes various microphones used for particular types of measurement. It is often desired to measure the sound pressure in a cavity or at a point where the presence of a normal microphone is prohibited because of its size or because its acoustic impedance would be low enough to modify the acoustic field it is desired to measure. In some cases, also, the temperature or sound pressure is too high for a normal unit to withstand. A small-bore probe tube made of suitable metal coupled to the front of a microphone may be used to conduct sound to the microphone proper, which can then work under reasonable ambient conditions. As these probes are generally used for measuring comparatively high sound levels in closed cavities, wave reflections can be substantially eliminated by packing the bore with porous absorbing material, the resultant increase in transmission loss in the probe being acceptable. Fig. 6(*a*) shows a moving-coil probe microphone and Fig. 6(*b*) shows its frequency response. The probe is a stainless-steel hypodermic tube connected to the diaphragm of the unit via a carefully designed acoustic transforming cavity.[9] The probe bore is damped with spaced tufts of absorbent. The small size (0·031 in) of the bore and the damping cause a loss of about 40 dB, but the high impedance looking into the open end of the probe makes the microphone suitable for sound-pressure measurements in artificial ear couplers, human ear canals, acoustic-impedance bridge couplers, etc. For work at high temperatures and pressures such as those in gas-turbine engine exhausts, various types of probe-tube condenser microphones have been developed.[10]

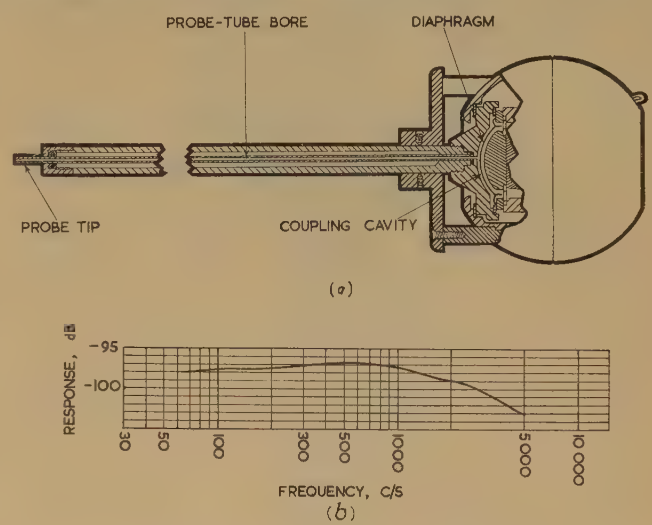


Fig. 6.—Pressure probe microphone.

(*a*) Cutaway view.
(*b*) Frequency response.
0 dB = 1 volt per dyne/cm².

Another type of microphone has been designed both as a probe and as a means of measuring free fields of high intensity. It consists of an expander bar of ammonium dihydrogen-phosphate piezo-electric crystal assembled in a small rigid-walled tube of about $\frac{3}{16}$ in diameter, the end being closed with a thin diaphragm cemented to the crystal face. The unit is omnidirectional to within 1 dB up to 8 kc/s, the frequency response being flat within 2 dB up to supersonic frequencies.[11] The sensitivity is necessarily rather low, and thus a carefully designed pre-amplifier must be used.

(3) DIRECTIONAL MICROPHONES

(3.1) Directivity Criteria

Directivity criteria apply to (*a*) bidirectional microphones whose response shows maximum sensitivity to the front and rear, (*b*) unidirectional microphones whose response is largely confined to the front and (*c*) highly-directional microphones whose response is confined to a limited angle at the front.

Various criteria have been defined which give quantitative

measures of directional properties. The *directivity factor* is the best known. This is defined in terms of the ratio of the efficiency for normal incident sound to the efficiency for sound of random incidence, and gives an indication of the average reduction in reverberant sound pick-up to be expected.

A *unidirectional factor* has also been proposed. This is defined in terms of the ratio of the average efficiency for all angles of sound incidence in the front hemisphere to the efficiency for all angles of sound incidence in the rear hemisphere. It gives a measure of ability to reject sounds from the rear. Fig. 7

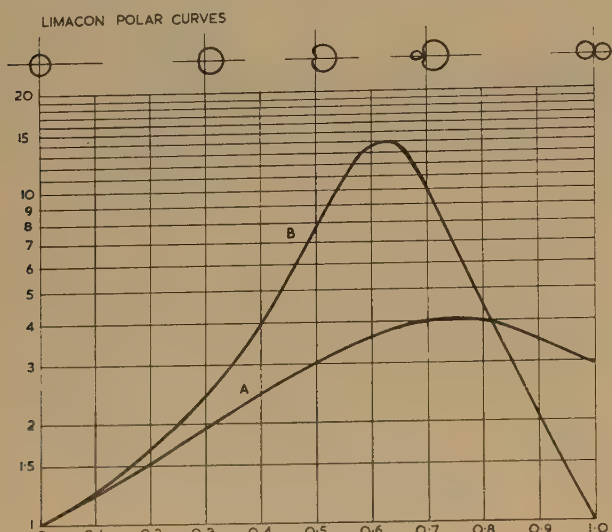


Fig. 7.—Limacon curves produced by pressure-gradient microphones.
K = Fractional contribution due to gradient component.
A = Directivity factor.
B = Unidirectional factor.

shows these factors associated with various limacon curves derived from the relationship

$$R(\theta) = (1 - K) + K \cos \theta$$

for values of K between 0 and 1. These curves are characteristic of first-order pressure-gradient-operated microphones incorporating various amounts of internal rear phase delay. Highly directional microphones of higher-order gradient or of 'wave' type may show much higher directivity or unidirectional factors. A normalizing magnitude factor and a sign convention which unfolds the limacon into front and rear lobes are invariably applied, so that readily intelligible polar curves are obtained.

The simplest first-order gradient microphone is the ribbon microphone. It consists of a ribbon suspended between two pole-pieces so that the front and rear directions have clear sound access. A bidirectional cosine-curve polar response results.

If an acoustic phase delay is introduced at the rear side, the response can be made unidirectional, the microphone being actuated, in effect, by a pressure and a gradient component. Two or more gradient units spaced apart along an axis with their outputs connected in electrical series opposition represent higher-order gradient combinations.[12]

(3.2) 'Wave'-Type Directional Microphones

'Wave'-type directional microphones depend primarily on the physical length of particular acoustic paths compared with the wavelength of sound. The simplest, and probably the oldest example is the reflector microphone, where an omnidirectional unit was mounted at the focal point of a paraboloid whose size was large compared with the wavelength of sound (e.g. 3 ft diameter was effective down to 800 c/s). One drawback to this microphone was the extreme narrowing of the polar response at high frequencies. This was compensated to some extent either by displacing the microphone slightly from the focal point of the paraboloid or by covering the outer parts of the paraboloid with a layer of acoustic absorbent, so that the higher frequencies were absorbed rather than reflected, thus reducing the effective size of the reflector at high frequencies. A further improvement on this system was proposed, using an acoustic lens and a horn.[13] Various types of 'line' microphones were developed as alternatives to the rather unwieldy reflector microphones. These line units consisted originally of a pressure-operated unit with a bundle of tubes of graded lengths coupling the sound field to the front of the diaphragm. A directional response was obtained because sounds incident along the axes of the tubes arrived in phase at the diaphragm, whereas sounds incident at other angles encountered differing path lengths through the staggered tubes so that a certain amount of out-of-phase pressure cancellation then occurred at the diaphragm.[23] It was later found possible to shorten the excessively long tubes needed to give directionality at low frequencies by introducing 'kinks' which provided extra acoustic delay in the tubes.

A still later example of a directional tubular 'line' microphone[14] operates in a similar way. It achieves a directional characteristic by means of a number of acoustically damped entries in the form of a line slit along a single tube 40 or 80 in long, the length being chosen according to the degree of directivity required and the discrimination desired at low frequencies. One end of the tube is coupled to the diaphragm of a moving-coil microphone unit (see Fig. 8). The presence of correctly

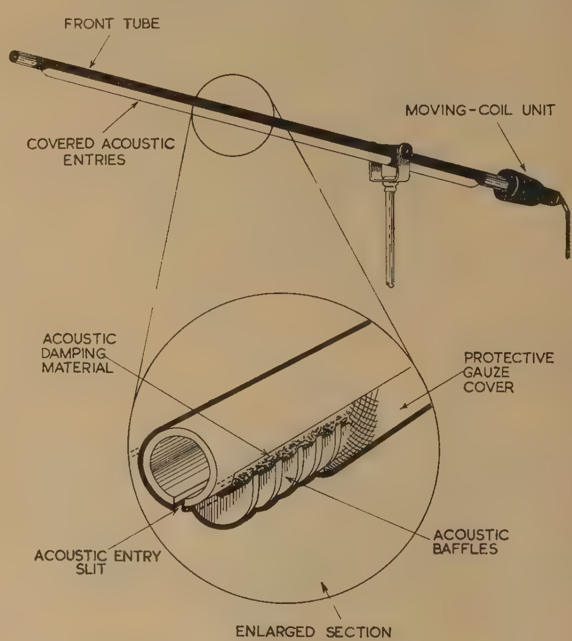


Fig. 8.—Tubular directional line microphone.

applied damping largely eliminates standing-wave effects along the tube, but it is necessary to increase the relative value of the acoustic resistance applied to entries nearer the diaphragm so that the sound-energy loss down the tube is compensated, the wave contributions from all entries arriving at the diaphragm with the same amplitude. The coupling of such a tube to a diaphragm via a coupling cavity results in a high-frequency loss which, in this case, is compensated by the addition of a serie

of small baffles mounted between the entries at 90° to the axis of the tube. These give a resonant-cavity gain at high frequencies. A further artifice is possible which reduces the effective length of the tube at high frequencies, thus minimizing the progressive narrowing of the polar response curve as the frequency is increased; this is achieved if a suitably increased acoustic mass reactance can be assigned to the entries near one end of the tube, thereby substantially eliminating their contribution at high frequencies. In effect, the entries form part of a distributed low-pass filter system with graded cut-off frequencies.

Column microphones have also been used to obtain directional effects. These consist of several microphone units arranged in a vertical column, their outputs being combined electrically in phase. A highly directional response is obtained in the vertical plane, owing to out-of-phase cancellation of the outputs of the various units for sound incident in the vertical plane at an angle to the normal.[15]

It is noteworthy that for the correct operation of a gradient unit its size must be reasonably small compared with the wavelength of sound, whilst the wave type of microphone must have dimensions at least comparable to the wavelength. The directional properties of the latter type fail at low frequencies where this condition does not hold.

(3.3) Studio Bidirectional Ribbon Microphones

Figs. 9(*a*) and (*b*) show a high-quality bidirectional gradient-operated ribbon microphone of fairly recent design and its response curve. The ribbon is made as short as possible in the vertical direction to preserve the high-frequency polar response, the ribbon being made as wide as possible in order to achieve a high efficiency.[16] The ribbon is a strip of pure aluminium leaf, 1 in long by $\frac{1}{4}$ in wide and about $\frac{1}{40}$ mil thick. It is corrugaged along its length in order to increase flexibility and to prevent any tendency for the edges to curl. The tension is adjusted to give a fundamental resonant frequency of 40–45 c/s. A ribbon of this kind has vibrational properties similar to those of a stretched string or an elastic bar, so that it is capable of vibration in many high modes in addition to the fundamental. It has been found that, for a very light ribbon of this kind, the resistive component of the air load, together with a slight additional coupled acoustic resistance due to open-weave metal gauzes mounted on either side of the ribbon, is able to produce critical damping of all the natural resonant modes of the ribbon. The motion of the ribbon is thus substantially aperiodic, and the response of the microphone is primarily determined by the dimensions and layout of the pole-pieces, magnet and case, etc. The ribbon is very light, its weight being about $0\cdot2\,\mathrm{mg/cm^2}$, and this, combined with the very effective acoustic damping, gives the microphone an outstandingly good transient response. A flat frequency response and high efficiency have also been achieved by careful proportioning of the pole-pieces and the magnet, the former being precision castings in vanadium Permendur, and the latter a high-energy-content magnet alloy with preferred directional properties. A comparatively high flux density of 6000 gauss is maintained across the $\frac{1}{4}$ in-wide ribbon. The driving force on the ribbon is approximately proportional to $d\cos\theta$, where d represents the effective distance corresponding to the sound pressure-gradient force, and θ is the angle of sound incidence relative to the normal. A flat response results if d is small compared with the wavelength of sound so that the pressure-gradient force is proportional to frequency, and if the ribbon acoustic or mechanical impedance is mass controlled. Several difficulties are encountered in practice. It is difficult to assign one simple value to d, and it is impossible to make the pole-piece cross-section uniform, the

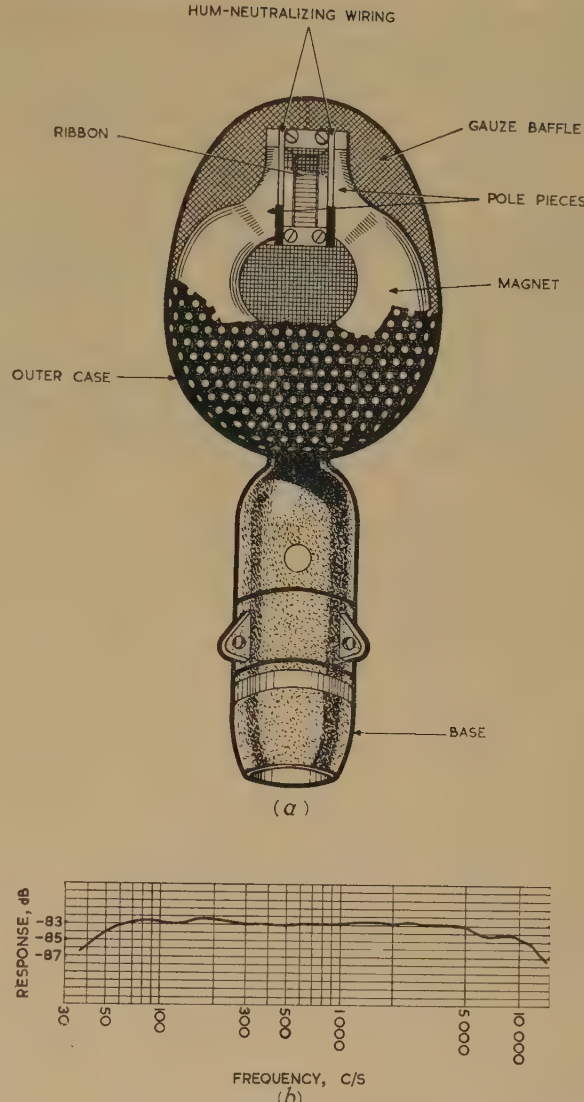


Fig. 9.—Studio ribbon microphone.
(*a*) Cutaway view.
(*b*) Free-field response.
0 dB = 1 volt per dyne/cm²; 0° sound incidence.

poles being enlarged at the bottom to meet the necessarily greater area of the magnet face. Thus d tends to have a larger value at the bottom of the poles than at the top. The larger distance gives a gradient driving force of larger magnitude but with a high-frequency cut-off at a lower frequency than is given by the smaller top distance. Such ribbon microphones are liable to have a step of several decibels in their frequency response at middle or high frequencies. In the present design, the pole-piece and magnet proportions are such that the effect is quite negligible. The response tends to fall at frequencies below about 150 c/s owing to the introduction of resistance rather than mass control, as a result of the critical damping of the natural resonances of the ribbon. A porous gauze baffle with a fairly low acoustic resistance is fitted round the pole-pieces and magnet window in order to compensate for this loss. The baffle performs this function because the pressure difference built up between its front and rear surfaces is a significant addition to the small low-frequency pressure differences produced across the ribbon surfaces by the pole-pieces, but is negligible compared with the

much larger pressure differences they produce at higher frequencies.[16] The natural tendency for the high frequencies to fall off as the distance d approaches a half-wavelength is compensated by a broad resonant-cavity gain due to the pole-piece chamfer and by a standing-wave reflection gain due to the spacing of the case from the ribbon plane, the acoustic impedance of the holes in the fluted upper part of the case having a preferred value. The total harmonic distortion of this microphone has a maximum value of less than 0·1% for the majority of frequencies and sound levels normally encountered in operation. The wiring of the ribbon is disposed so as to neutralize electromagnetic hum pick-up, and the toroidally wound transformer in the stem is fitted with a Permalloy-C shield.

The extremely low mechanical impedance of the ribbon renders these microphones likely to produce 'rumble' from structure-borne vibrations or the relative movement of the air caused either by draughts or by movement of the microphone on a boom mounting. Suitable shock mountings and wind shields have been designed to overcome these objections.

(3.4) Unidirectional Gradient Microphones

Fig. 10 shows diagrammatically the construction, analogous electrical circuit and frequency response of a moving-coil first-order-gradient unidirectional microphone. The essential feature which distinguishes this type of unit from a simple bidirectional gradient microphone is the provision of acoustic phase-shifting elements coupling the rear side of the diaphragm to the outside air. The vector diagrams of Fig. 11 show how the acoustic phase-shift principle is applied to a gradient unit to give a unidirectional response. The rotating vector p_1 represents the alternating acoustic pressure at the front opening of the diaphragm. Similarly, p_2 represents the pressure at the rear openings which are part of the rear phase-shifting system, and p_3 the pressure immediately behind the diaphragm, produced by the action of p_2 on the phase-shifting network. The net pressure available for driving the diaphragm is the vector difference between p_1 and p_3. If we take p_2 as a reference vector and examine the difference vector $p_1 \sim p_3$ as the angle of sound incidence is varied from the normal, 0°, we see that for 0° incidence, $p_1 \sim p_3$ is a maximum and is approximately twice the magnitude of $p_1 \sim p_2$. For 90° incidence, $p_1 \sim p_3$ is reduced by about 6 dB, whilst for 180° incidence $p_1 \sim p_3$ is substantially zero. The main requirements of the rear phase-shift system are that it should introduce a phase shift proportional to frequency and equal or proportional to that encountered by the sound wave in traversing the external distance d between the front and rear openings on the microphone. The polar responses which can be obtained are of the shapes derived from the limacon:

$$R(\theta) = (1 - K) + K\cos\theta \quad . \quad . \quad . \quad . \quad (1)$$

the best known of which is the cardioid

$$R(\theta) = \tfrac{1}{2}(1 + \cos\theta) \quad . \quad . \quad . \quad . \quad . \quad (2)$$

which is obtained when the internal and external phase angles are equal. The theory of these 'phase-shift' unidirectional microphones is worked out more completely in Section 8.2. It is seen there that, in the microphone considered above, the main-case side-opening holes M_5, R_5 must be substantially resistive and must feed a cavity, S_3, behind the diaphragm whose value in relation to the resistance of the openings is determined by the distance d and the particular polar curve desired. The other requirement to be met is that the moving system of this microphone must be substantially mass controlled if the gradient force, which rises with frequency, is to give it a flat frequency response. Both these requirements are difficult to meet over a wide frequency range, and in most practical microphones special means have to be adopted to obtain the desired results at both extremes of the frequency scale. The method adopted in this microphone to obtain more effective mass control at low frequencies is to couple acoustic masses, such as the air in the pipes M_4, M_6 and the element M_3 in Fig. 10(*b*) to the diaphragm by means of the small cavities S_1, S_2 behind it. The opening of M_6 to the air at a greater distance from the front gives rise to an enhanced gradient force operative at low frequencies, M_6 and R_4, etc., giving the appropriate phase shift. The resonator M_4, S_4 is tuned to a correspondingly low frequency and is damped by R_4, the governing factor at resonance. Above about 4 kc/s the action of the phase-shifting elements becomes complicated owing to the onset of wave effects caused by the wavelength of sound becoming comparable with the dimensions of the elements concerned. It is therefore arranged that the unit shall change gradually to a pressure-operated resistance-controlled device at high frequencies

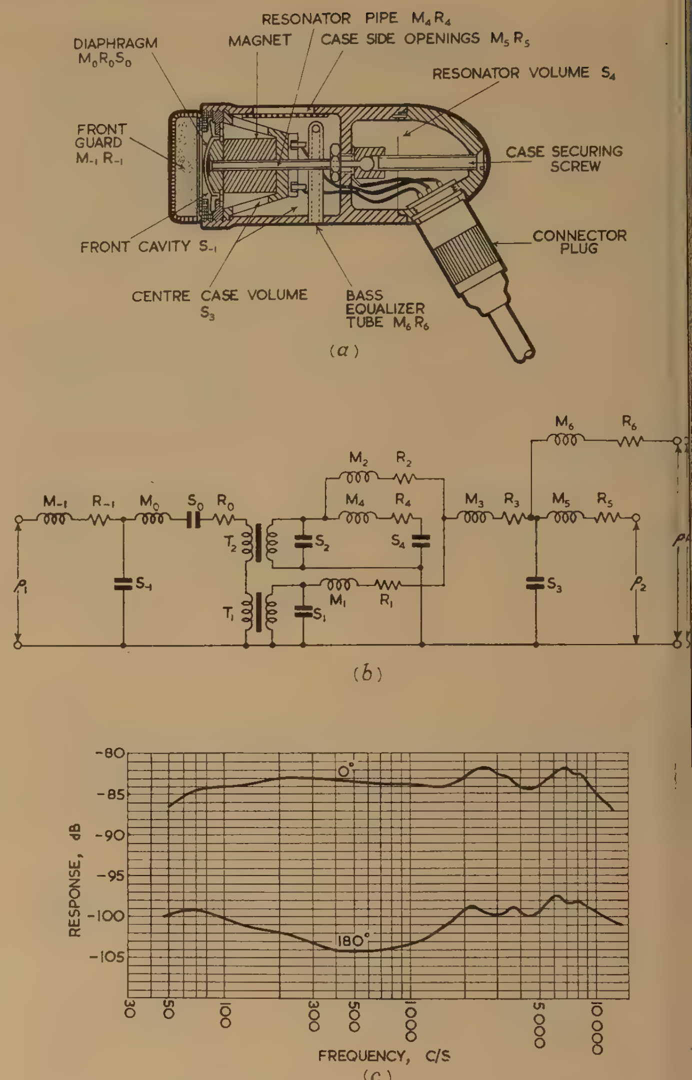


Fig. 10.—Unidirectional moving-coil microphone.
(*a*) Cross-sectional view.
(*b*) Analogous electrical circuit.
p_1 = Acoustic pressure on front openings.
p_2 = Acoustic pressure on case side openings.
p_2' = Acoustic pressure on bass equalizer pipe opening.
T_1 and T_2 are ideal transformers.
(*c*) Free-field response.
0 dB = 1 volt per dyne/cm².

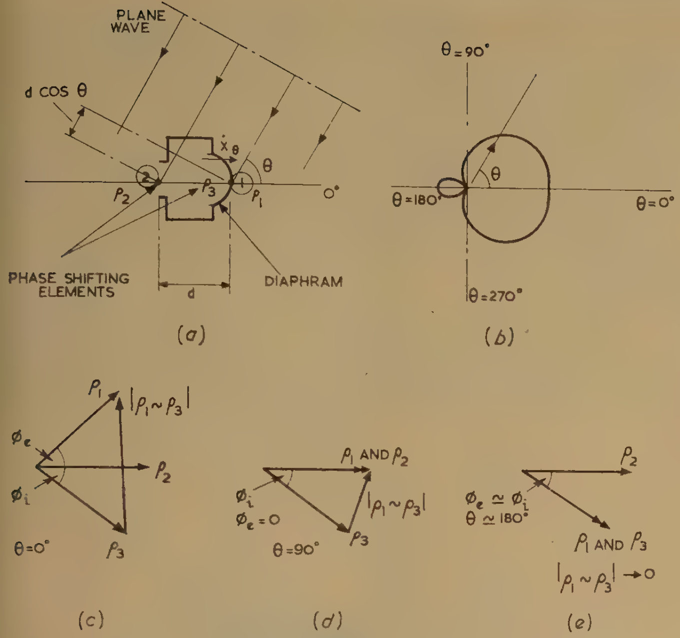


Fig. 11.—Relationships involved in the operation of a gradient unidirectional microphone.

(*a*) Diaphragm of microphone showing external and internal sound pressure due to sound incident at θ° to normal.
(*b*) Typical unidirectional polar response obtained.
(*c*) Sound-pressure vectors for $\theta = 0°$.
(*d*) Sound-pressure vectors for $\theta = 90°$.
(*e*) Sound-pressure vectors for $\theta = 180°$.

$$\phi_e = \text{Phase shift in sound wave between external openings} = \frac{\omega d}{v}.$$

$$\phi_i = \text{Phase shift in microphone internal elements} = \frac{\omega d}{v}\frac{(1-K)}{K}.$$

the unidirectional response being obtained by diffraction effects from the casing.[19, 20]

Ribbon,[26] crystal[21] and condenser[5, 27, 28] units have also been designed to function as phase-shift unidirectional microphones. In all cases the phase-shift elements, the external distance and the characteristics of the moving system have to be co-ordinated in order to meet the frequency-response and unidirectivity requirements.

Another development is a unidirectional microphone in which a substantially resistance-controlled moving-coil system is used, a constant force being derived by means of a series of openings at different distances from the diaphragm, the paths connected to the various openings being arranged to operate over appropriate frequency bands.[22]

The condenser microphones operate with constant amplitude over the frequency range. This is achieved with a gradient unit using a resistance-controlled diaphragm backed by an acoustic phase-shifting network with a rear opening. For a constant unidirectional response at all frequencies, this type of microphone probably affords the most elegant solution, since it is possible to make the unit small enough to minimize wave effects and thus largely to avoid compromises involving the use of case diffraction effects to obtain unidirectivity at high frequencies.

By means of remote switching of the polarizing voltage for the main and an auxiliary diaphragm, fitted over the rear opening, it is possible to make the polar response take up the various curves of the limacon $R(\theta) = (1 - K) + K\cos\theta$, and thus to have available various polar curves between omnidirectional and unidirectional.[5] This occurs, in effect, because the proportion, K, of the output due to the acoustic gradient component is varied. The general principle is easy to understand when one of the earlier types of unidirectional microphone is studied. This consists of a moving-coil pressure unit and a ribbon bidirectional gradient unit with their outputs electrically combined in series. When the output of the gradient unit is varied by means of taps on its matching transformer, some of the various polar curves of the limacon are obtained.[23]

(4) CLOSE-TALKING AND NOISE-REDUCING MICROPHONES

High-quality versions of close-talking and noise-reducing microphones have been designed so that broadcast commentaries, etc., can be given under conditions where the level of background noise requires that the microphone be used very close to the mouth. This involves special problems in securing natural-sounding speech reproduction, and any noise-reducing properties which can be conferred on the microphones are valuable. In studying close-range speech, account must be taken of the four sound sources concerned, namely mouth, nose, throat and chest.

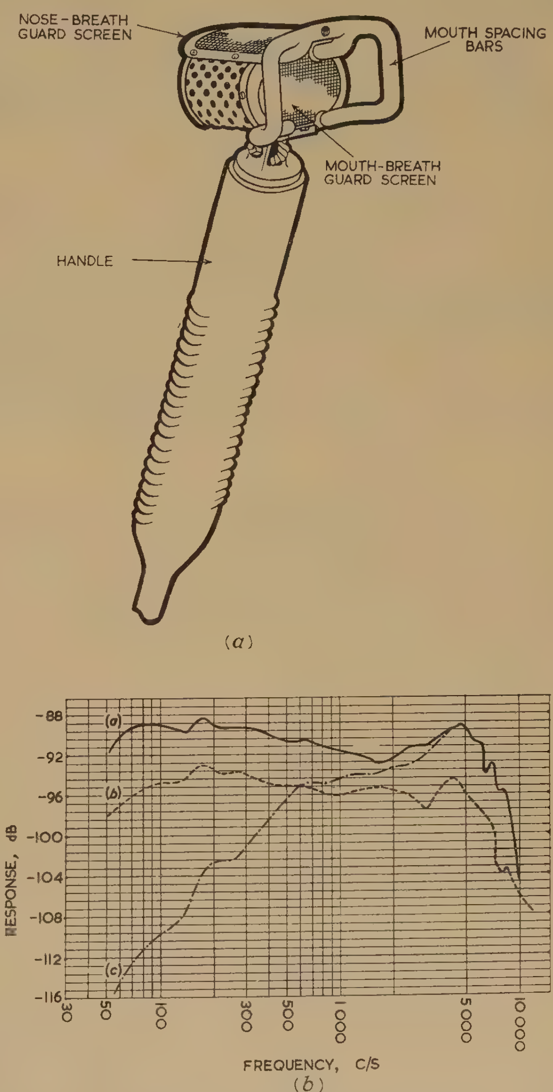


Fig. 12.—Close-talking microphones with noise-suppressing properties.

(*a*) Diagrammatic view.
(*b*) Frequency responses.

——— Response to a mouth source at 2 in distance.
– – – – Response to a nose source corresponding to the full-line curve
—·— Response to a free plane wave.

0 dB = 1 volt per dyne/cm²

The last two, being predominantly low-frequency sources, are usually of secondary importance, but a balanced reproduction must be obtained from the nose and mouth sources. Other problems are concerned with breath noises and the effects of explosive speech sounds, for example certain consonants such as P. In addition, the frequency spectrum of sound energy in close-range speech differs from that in speech at normal distances.

Fig. 12(*a*) shows a close-talking noise-reducing commentator's microphone which has been developed for broadcasting. It is a gradient ribbon unit with a high degree of acoustic damping, fitted into a special case equipped with nose and mouth breath shields; the talking distance is accurately controlled by means of positioning bars which contact the talker's upper lip. Noise reduction occurs, first, because the bidirectional cosine polar response affords a theoretical reduction of about 66% in the pick-up of random noise energy, and secondly, because the proximity effect of a first-order gradient microphone of these proportions is such that, in order to have a flat frequency response for the correct talking distance of about 2 in between the lips and the ribbon, the free-field frequency response has to show a fall of about 30 dB between 50 c/s and 3 kc/s, thus greatly attenuating low- and middle-frequency background noise.[24] Fig. 12(*b*) shows these frequency responses. The rising response at about 5 kc/s compensates known high-frequency close-talking losses.

Other features of the microphone are the use of stainless-steel woven mesh for breath shields, which avoids the use of materials that might become sodden or unhygienic under prolonged close-talking use, and the provision of a high degree of neutralization and shielding against electromagnetic induction over a frequency range of 50 c/s to over 10 kc/s. This is necessary in order to avoid induction from sources such as television monitor sets. The speech quality is good, mainly owing to the correct balance being obtained between the nose and mouth tones; any deficiency of the former gives a 'stuffy' and unnatural speech quality.

(5) WIND-SHIELDING AND WEATHER PROTECTION OF MICROPHONES

Microphones used out of doors may have to be fitted with some means of protection against wind and rain, and it is desirable that the performance and use of the microphone should not be adversely affected by any wind or rain shield.

It is thought that the causes of wind noise can be considered as falling into three categories:

(*a*) Shock excitation of the diaphragm, owing to fluctuating pressure wavefronts of the wind.

(*b*) 'Edge tones' at characteristic frequencies dependent for their frequency and amplitude on the spacing and size of discontinuities such as edges, bars or irregularities on the microphone body.

(*c*) Tones due to eddies or turbulence generated by the flow of wind around the microphone case as a whole. The frequency of these tones is inversely proportional to the obstacle size, being in the region of 50–200 c/s, for microphones of normal size.

The usual subjective effects of wind noise are heavy low-frequency rumbling sounds, together with higher-frequency impulsive noises and hissing sounds.

An ideal wind shield needs to offer a high filtration loss to the substantially unidirectional air currents of the wind, whilst leaving audio-frequency sounds relatively unaffected. This can be achieved by a comparatively large shield, preferably spherical, of about 9–12 in diameter, the outer surface being of closely woven sheer fabric, very fine wire meshes, or porous materials possessing suitable characteristics. This type of wind shield still gives rise to turbulent noises, but its size is such that these are mainly confined to the sub-audible frequency range. The overall subjective reduction in wind noise can be 20 dB or more.

Large wind shields of this type are often not acceptable because they are bulky to handle and obtrusive in appearance. It is therefore necessary to devise a smaller means of producing a fair degree of immunity to wind. This can be done at the expense of a small loss in extreme high-frequency response by a closely fitting shield of appropriate porous or foamed plastic material which encloses the front or other openings of the microphone.

In the design of the microphone, a suitable amount of acoustic damping of the moving system and a case having the smoothest possible surface help to reduce disturbances in the first two categories.

Porous or foam materials usually require some protection in the form of an enclosing shield of fine woven wire mesh, which may be treated with a water repellent such as a silicone varnish, in order to prevent the inner porous material from becoming sodden. If water is allowed to collect on these parts or on other parts involved in acoustic transmission within the microphone, a considerable loss of sensitivity, particularly at the higher frequencies, is likely to occur.

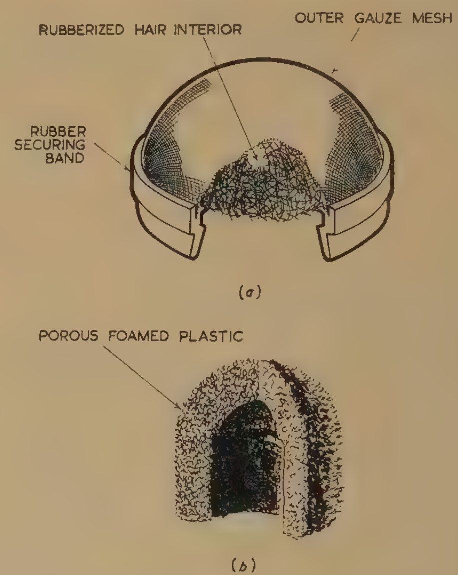


Fig. 13.—Microphone wind shields.
(*a*) Rubberized hair protected by stainless-steel woven mesh.
(*b*) Silicon-treated foamed plastic.

A wind shield of the closely fitting kind cannot greatly alter the wind-turbulence tones. A well-designed shield of this type, used in conjunction with some electrical bass cut-off below 200 c/s, enables an overall attenuation of wind noise up to 20 dB to be achieved, and allows good speech to be transmitted in fresh or strong winds of up to 40 m.p.h. The effect on frequency response may be confined to a loss of 2–3 dB at 10 kc/s. Fig. 13 shows unobtrusive wind shields of the above type.

(6) ACKNOWLEDGMENTS

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(8) APPENDICES

(8.1) Derivation of the Analogous Electrical Circuit of an Omnidirectional Moving-Coil Microphone

First a functional diagram is drawn from a knowledge of the mechanical construction of the microphone. The various parts acting as masses are identified by reason of their ability to store kinetic energy, and their vibrational mode displacements are chosen as independent co-ordinates. Fig. 14 gives the functional

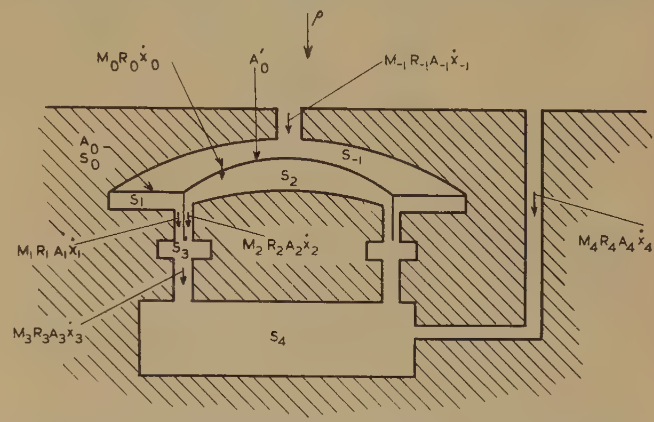

Fig. 14.—Functional diagram used in the derivation of the analogous electrical circuit of a microphone.

diagram of the microphones shown in Figs. 2(*a*) and (*b*). The various constants in the form of lumped masses, resistances and stiffnesses are identified, and the appropriate areas and linear velocities are designated.

From Lagrange's equation,[25]

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{x}_n}\right) + \left(\frac{\partial F}{\partial \dot{x}_n}\right) + \left(\frac{\partial V}{\partial x_n}\right) = e_n(t) \qquad (3)$$

where $e_n(t)$ = Applied force for each co-ordinate.
$\dot{x}_n$ = Linear co-ordinate velocities.
x_n = Linear co-ordinate displacements.
$n = -1, 0, 1, \ldots$
T = Kinetic-energy function.
F = Dissipational-energy function.
V = Potential-energy function.

The kinetic-energy function is derived by inspection of Fig. 14:

$$T = \tfrac{1}{2}M_{-1}A_{-1}^2\dot{x}_{-1}^2 + \tfrac{1}{2}M_0(A_0 + A_0')^2\dot{x}_0^2 + \tfrac{1}{2}M_1A_1^2\dot{x}_1^2 + \tfrac{1}{2}M_2A_2^2\dot{x}_2^2 + \tfrac{1}{2}M_3A_3^2\dot{x}_3^2 + \tfrac{1}{2}M_4A_4^2\dot{x}_4^2 \qquad (4)$$

Similarly, for the dissipational-energy function,

$$F = \tfrac{1}{2}R_{-1}A_{-1}^2\dot{x}_{-1}^2 + \tfrac{1}{2}R_0(A_0 + A_0')^2\dot{x}_0^2 + \tfrac{1}{2}R_1A_1^2\dot{x}_1^2 + \tfrac{1}{2}R_2A_2^2\dot{x}_2^2 + \tfrac{1}{2}R_3A_3^2\dot{x}_3^2 + \tfrac{1}{2}R_4A_4^2\dot{x}_4^2 \qquad (5)$$

and for the potential-energy function,

$$V = \tfrac{1}{2}S_{-1}[A_{-1}x_{-1} - (A_0 + A_0')x_0]^2 + \tfrac{1}{2}S_0(A_0 + A_0')^2x_0^2 + \tfrac{1}{2}S_1(A_0x_0 - A_1x_1)^2 + \tfrac{1}{2}S_2(A_0'x_0 - A_2x_2)^2 + \tfrac{1}{2}S_3(A_1x_1 + A_2x_2 - A_3x_3)^2 + \tfrac{1}{2}S_4(A_3x_3 + A_2x_4)^2 \qquad (6)$$

The constants M_{-1}, M_0, $M_1 \ldots$, R_{-1}, R_0, $R_1 \ldots$ and S_{-1}, S_0, $S_1 \ldots$ are assumed to be in analogous acoustic impedance units and thus are multiplied by the square of the area concerned in order to give the equivalent mechanical constants.

Taking the x_{-1} co-ordinate, differentiating eqns. (4), (5) and

(6) with respect to $\dot{x}_{-1}$, x_{-1} and t as indicated and substituting in eqn. (3), we obtain

$$j\omega M_{-1}A_{-1}\dot{x}_{-1} + R_{-1}A_{-1}\dot{x}_{-1} + \frac{S_{-1}}{j\omega}[A_{-1}\dot{x}_{-1} - (A_0 + A_0')\dot{x}_0] = p \quad (7)$$

Similarly, taking the x_0 co-ordinate, we obtain

$$j\omega M_0(A_0 + A_0')\dot{x}_0 + R_0(A_0 + A_0')\dot{x}_0 + \frac{S_{-1}}{j\omega}[(A_0 + A_0')\dot{x}_0 - A_{-1}\dot{x}_{-1}] + \frac{S_0}{j\omega}(A_0 + A_0')\dot{x}_0 + \frac{S_1}{j\omega}\frac{A_0}{A_0 + A_0'}(A_0\dot{x}_0 - A_1\dot{x}_1) + \frac{S_2}{j\omega}\frac{A_0'}{A_0 + A_0'}(A_0'\dot{x}_0 - A_2\dot{x}_2) = 0 \quad . \quad . \quad . \quad . \quad . \quad (8)$$

Taking the x_1 co-ordinate,

$$\frac{S_1}{j\omega}(A_1\dot{x}_1 - A_0\dot{x}_0) + j\omega M_1A_1\dot{x}_1 + R_1A_1\dot{x}_1 + \frac{S_3}{j\omega}(A_1\dot{x}_1 + A_2\dot{x}_2 - A_3\dot{x}_3) = 0 \quad (9)$$

Taking the x_2 co-ordinate,

$$\frac{S_2}{i\omega}(A_2\dot{x}_2 - A_0'\dot{x}_0) + j\omega M_2A_2\dot{x}_2 + R_2A_2\dot{x}_2 + \frac{S_3}{j\omega}(A_2\dot{x}_2 + A_1\dot{x}_1 - A_3\dot{x}_3) = 0 \quad (10)$$

Taking the x_3 co-ordinate,

$$j\omega M_3A_3\dot{x}_3 + R_3A_3\dot{x}_3 + \frac{S_3}{j\omega}(A_3\dot{x}_3 - A_1\dot{x}_1 - A_2\dot{x}_2) + \frac{S_4}{j\omega}(A_3\dot{x}_3 + A_4\dot{x}_4) = 0 \quad (11)$$

Taking the x_4 co-ordinate,

$$j\omega M_4A_4\dot{x}_4 + R_4A_4\dot{x}_4 + \frac{S_4}{j\omega}(A_4\dot{x}_4 + A_3\dot{x}_3) = p \quad . \quad (12)$$

Eqns. (7)–(12), inclusive, are seen to be the mesh equations of the analogous electrical circuit given in Fig. 3. Eqn. (8) shows that the two separately acoustically loaded areas, represented by the rear surface of the diaphragm dome and the rear surface of the diaphragm surround, lead to the two ideal transformers, T_1 and T_2, which have turns ratios of $A_0/(A_0 + A_0')$ and $A_0'/(A_0 + A_0')$, respectively.

The input voltage applied to the network represents the sound pressure applied to the microphone front openings, whilst the current through M_0, the inductance representing the diaphragm and speech coil moving mass, is proportional to the e.m.f. generated or the response of the microphone. In this analysis, wave effects are assumed to be absent and thus it is legitimate to assume that the same sound pressure is applied to the microphone front openings M_{-1} and to the rear leak tube M_4 openings, without any phase displacement, and also that the microphone main case volume can be represented by a single stiffness S_4. In practice, the fact that the leak tube is only operative at frequencies below 100 c/s ($\lambda > 10$ft), and the employment of distributed damping in the microphone case volume, mean that the approximation is valid.

It will be noted that the analogous circuit normally gives the pressure response rather than the free-field response of the microphone, as the effects of diffraction around the case cause a modification of the constant-input-voltage condition for the circuit.

(8.2) Theory of Single-Unit Gradient Unidirectional Microphones

In general terms, the unit must have sound openings 1 and 2 separated by a distance d small compared with the sound wavelength. The internal arrangements of the microphone can be considered to be a passive acoustic-mechanical network, which as a result of sound pressure incident on the two openings and mechano-electric transduction, can cause an electric current to flow into an electrical termination.[21] Fig. 11(a) illustrates the arrangement. The motion of the transducer element can be considered as

$$\dot{x}_\theta = p_1\alpha - p_2\beta \quad . \quad . \quad . \quad . \quad . \quad (13)$$

where p_1 and p_2 are the pressures at the openings 1 and 2 resulting from the wave incident at an angle θ, and α and β are coefficients involving the network elements.

From eqn. (13),

$$\dot{x}_\theta = \alpha\left(p_1 - \frac{\beta}{\alpha}p_2\right) \quad . \quad . \quad . \quad . \quad . \quad (14)$$

For unidirectional (cardioid) response, $\dot{x}_\theta = 0$ for $\theta = 180°$.

Then
$$\frac{\beta}{\alpha} = \frac{p_1}{p_2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (15)$$

To a first order, we can assume that $|p_1| = |p_2|$ and that

$$p_1 = p_2\varepsilon^{j\omega\frac{d}{v}\cos\theta}$$

Substituting this in eqn. (15),

$$\frac{\beta}{\alpha} = \varepsilon^{-j\omega\frac{d}{v}}, \text{ as } \cos\theta = -1 \text{ for } \theta = 180°$$

and substituting this expression into eqn. (14),

$$\dot{x}_\theta = \alpha\left(p_1 - p_2\varepsilon^{-j\omega\frac{d}{v}}\right) = \alpha p_2\left(\varepsilon^{j\omega\frac{d}{v}\cos\theta} - \varepsilon^{-j\omega\frac{d}{v}}\right)$$

Therefore

$$\dot{x}_\theta = \alpha p_2\left[\left(1 + j\omega\frac{d}{v}\cos\theta\ldots\right) - \left(1 - j\omega\frac{d}{v}\ldots\right)\right]$$

and
$$\dot{x}_\theta = \alpha p_2 j\omega\frac{d}{v}(1 + \cos\theta) \quad . \quad . \quad . \quad . \quad (16)$$

provided that $d \ll \lambda$, so that $\omega d/v$ is small and thus the higher terms of the exponential expansions may be neglected. Also

$$p = |p_1| \simeq |p_2|$$

This analysis shows that a cardioid polar response is obtained, provided that the network produces a phase shift proportional to frequency. It also shows that, with a constant-velocity electromechanical transducer, the magnitude of the admittance coefficient, α, of the network including the moving part of the transducer element must be inversely proportional to frequency, i.e. it must be mass controlled [see eqn. (16)].

Thus, there are two conditions which the network as a whole must fulfil if a satisfactory unidirectional microphone is to result.

From the above theory and the vector diagrams of Fig. 11, it is seen that the rear acoustic network must provide a phase shift which is proportional to frequency and is either equal to, or a definite fraction of, the external gradient phase shift, according to whether a cardioid or hyper-cardioid polar response is desired. When the analogous electrical circuit of such a microphone is studied, values for the phase-shifting elements can be obtained. In an actual microphone there may be additional phase shift owing to the finite internal distances and the effects of wave propagation inside the case, etc., and so values calculated from

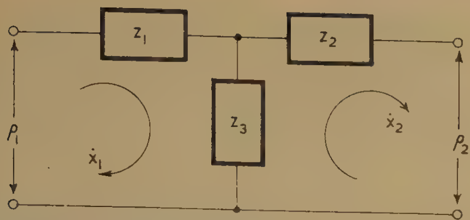

Fig. 15.—Simplified analogous circuit for a 'phase-shift' unidirectional microphone.

the lumped-constant analogous circuit may be inaccurate, but, as before, this type of circuit can still be very helpful.

Fig. 15 shows a simplified analogous circuit in the form of three impedances: Z_1 representing the diaphragm system, Z_3 the impedance of the elements backing the diaphragm, and Z_2 the impedance of the rear opening holes. Z_2 and Z_3 are thus the phase-shifting elements. Writing down the network equations,

$$p_1 = \dot{x}_1(Z_1 + Z_3) - \dot{x}_2 Z_3$$

$$-p_2 = -\dot{x}_1 Z_3 + \dot{x}_2(Z_2 + Z_3)$$

Therefore $$\dot{x}_1 = \frac{p_1(Z_2 + Z_3) - p_2 Z_3}{Z_1 Z_2 + Z_1 Z_3 + Z_2 Z_3} \quad . \quad . \quad . \quad (17)$$

But $p_2 \simeq p_1\left(1 - j\frac{\omega}{v} d\cos\theta\right)$, if $\frac{\omega d}{v}$ is small. where v is the velocity of propagation of sound.

Substituting in eqn. (17),

$$\dot{x}_1 = \frac{p_1\left[Z_2 + Z_3\left(j\frac{\omega d}{v}\cos\theta\right)\right]}{Z_1 Z_2 + Z_1 Z_3 + Z_2 Z_3}$$

If $\dfrac{j\frac{\omega d}{v}Z_3}{Z_2} = 1$ is substituted in the above equation,

$$\dot{x}_1 = \frac{P_1 Z_2(1 + \cos\theta)}{Z_1 Z_2 + Z_1 Z_3 + Z_2 Z_3} \quad . \quad . \quad . \quad (18)$$

This represents a cardioid polar response.

If Z_2 is a resistance R_2 and Z_3 is a compliance C_3,

$$\frac{j\omega d}{j\omega v C_3 R_2} = 1$$

Therefore $$\frac{d}{vC_3R_2} = 1 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (19)$$

If Z_2 is a mass M_2 and Z_3 is a resistance R_3,

$$\frac{j\omega d R_3}{j\omega v M_2} = 1$$

Therefore $$\frac{dR_3}{vM_2} = 1 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (20)$$

Eqns. (19) and (20) give the conditions for a cardioid response using the two basic types of phase-shifting elements. The first arrangement is used in some condenser cardioid microphones,[5, 27] and the second has been used in some moving-coil and ribbon unidirectional microphones.[20, 26]

The microphone illustrated in Fig. 10, uses, in effect, both types, the RC type represented by R_5 and S_3 operating over most of the frequency range, but with the R/M configuration represented by the pipe M_6 and R_4, etc., helping at extreme low frequencies. If a response other than cardioid is desired, such as the limacon for $K = 0{\cdot}63$ in Fig. 7, the required internal phase shift is reduced by the factor $(1 - K)/K$ or $(1 - 0{\cdot}63)/0{\cdot}63 = 0{\cdot}59$. This particular curve is seen to give the highest value of the unidirectional factor for a simple first-order-gradient unit of this type.

More elaborate phase-shifting networks have been described which have the object of achieving the frontal response of the hypercardioid without the rear lobe, thus giving an even higher unidirectional factor. The principle employed is to couple an additional sound opening, connected to a damped cavity, with the usual mass-reactive rear-entry holes.[26]

DISCUSSION BEFORE THE ELECTRONICS AND COMMUNICATIONS SECTION, 23RD MARCH, 1959

Mr. H. A. M. Clark: Perhaps the main feature in connection with microphones in stereophonic pick-up is their polar diagram. There is an increasing interest in types of stereophonic sound pick-up which utilize the polar diagrams of microphones as the origin of the stereophonic effect. Consequently, the point to which the author refers in Section 1.1(*b*) becomes of paramount importance, for otherwise we shall have a directional characteristic of the microphone varying with frequency, which is, in general, a highly undesirable feature.

Fig. A shows a ribbon microphone, constructed about 24 years ago. The two ribbons are in quadrature and the corrugations referred to in the paper are there in an elementary form. It was considered desirable to put the two ribbons in the same horizontal plane instead of one above the other, which resulted in the low-efficiency magnetic pole system which is seen mounted on the electromagnet. The ribbons were only 2½cm long in order to keep the vertical polar diagram good. This microphone had a most constant polar diagram, more so in fact than any directional microphone which was available at that time. The results obtained with it were satisfactory, and in some respects are difficult to improve upon to-day.

Diagrams other than the cosine diagram are sometimes required for stereophonic purposes, and this has caused the electrostatic microphone, which can have a polar diagram of either the cardioid or cosine type, to return to commercial use. With electrostatic microphones, in which the diameter can be reduced to about ½in, it is easier to secure the required polar diagram at high frequencies than in those using moving-coil elements, but it is more difficult to obtain the required diagram at low frequencies; the crevices between the circles of the cosine, and the crevices at the rear of the cardioid tend to be filled in. On the whole it seems to be easier to make the cosine type of microphone with a more consistent polar diagram than a cardioid.

Fig. B shows the polar response that can be obtained with a small electrostatic microphone of modern construction having two elements, one with a cardioid and the other with a cosine diagram with its axis at 90° to the first. Measurements are shown for 30, 60, 200 and 1 000 c/s. The curves match the true cardioid and cosine diagrams comparatively well. It is found much more difficult to match a pair of microphones with cardioid diagrams than with cosine diagrams, but by using one of each type, and by summing and differencing the two, one can obtain two diagrams which are very similar to cardioids and identical in shape.

Fig. C shows the resulting curves from 30 c/s to 1 kc/s, and Fig. D those from 1 to 10 kc/s.

In the paper, frequency response curves for various micro-

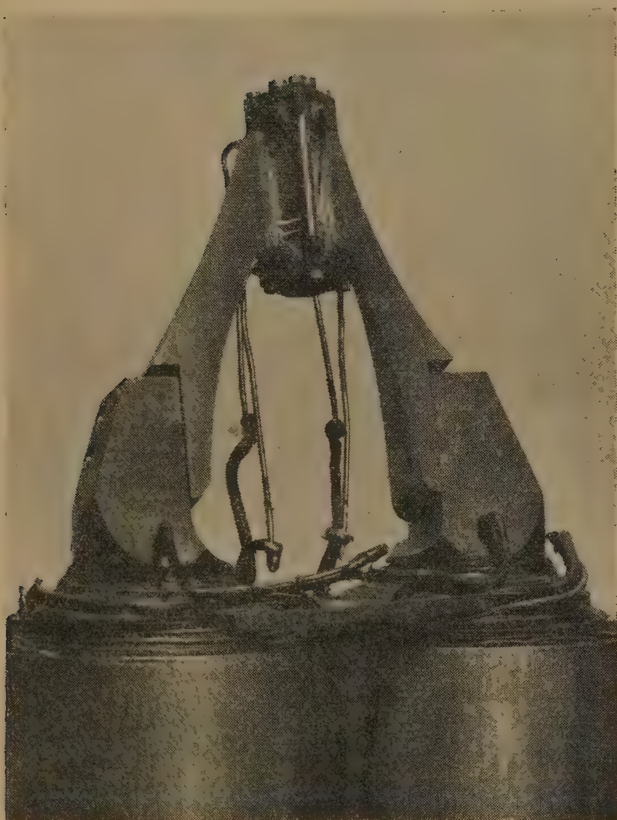

Fig. A.—Early type of ribbon microphone used for stereophonic recording.

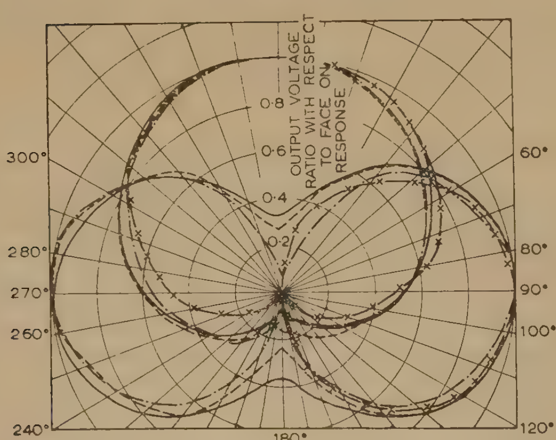

Fig. B.—Horizontal polar characteristics of cardioid-cosine microphones.

—— 30 c/s. —·— 200 c/s.
---- 60 c/s. —×— 1 kc/s.

phones are given plotted in decibels relative to 1 volt per dyne/cm^2. These give no idea of the output of the microphone unless the intended load impedance is also quoted. Would not a scale of decibels above 1 mW into a specified load give a quick comparison of sensitivities?

At the high-frequency end of the audio range an increase in the frequency bandwidth, particularly of the loudspeakers, accentuates many faults in the system. The loudspeakers and amplifiers may be measured very carefully, without finding any

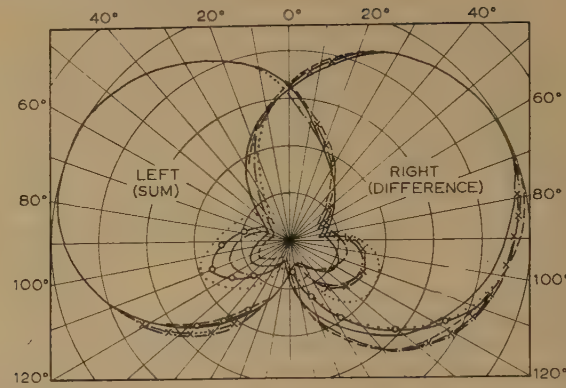

Fig. C.—Sum-and-difference polar diagrams of cardioid-cosine electrostatic microphones.

—— 30 c/s. —×— 200 c/s.
---- 60 c/s. ······ 300 c/s.
—·— 120 c/s. —○— 450 c/s.

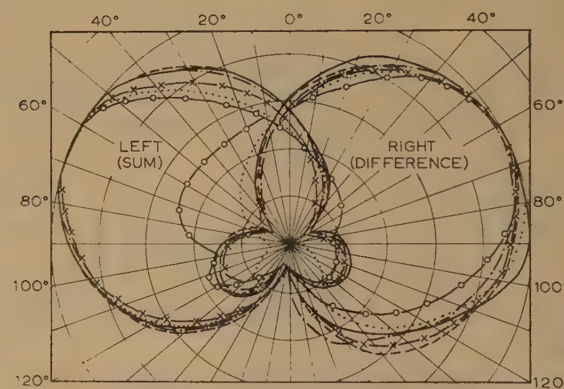

Fig. D.—Sum-and-difference polar diagrams of cardioid-cosine electrostatic microphones.

—— 1 kc/s. —×— 6 kc/s.
---- 2 kc/s. ······ 8 kc/s.
—·— 4 kc/s. —○— 10 kc/s.

reason for this distortion. Can the author give any details on th performance of microphones from the linearity point of view a high frequencies? Is the assumption that all movements ar small enough to avoid non-linear distortion correct? The high frequency content of many instruments is very high, and, fo special effects, microphones are sometimes placed close to th instruments, where the sound level at high frequencies may b surprisingly high. But having determined that level, it is ra that the manufacturer can indicate what the intermodulatio distortion may be there.

Mr. D. W. Robinson: An important feature of instrumen microphones is the limitation imposed by thermal noise in th effective series-resistance component of the electrical impedanc or by amplifier noise. Absolute values of the equivalent sounc pressure threshold depend primarily on acousto-electric efficienc What is the author's view on the improvements possible in th direction with electrodynamic or electrostatic microphones The threshold, with practical circuits, is of about the same ord for modern microphones of both classes, and usually around c just above 0 dB relative to 0·0002 dyne/cm^2 per octave ban Owing to the dissimilar electrical impedances, however, th equivalent threshold increases with frequency for the electr dynamic type but falls for the electrostatic, which complicat the comparison.

It would be useful if a calibrating resistor were incorporate

between the low-potential microphone terminal and earth during manufacture, which would greatly simplify maintaining the calibration of associated measuring equipment.

The probe microphones mentioned in the paper can, with ingenuity, be made to exhibit fairly flat frequency responses. Nevertheless, pressure at the orifice is related to diaphragm motion by complicated expressions involving frequency through the arguments of Bessel functions. I have experimented with the simple principle of a 'semi-infinite' tube, interrupted, without discontinuity of cross-section, by a radially polarized hollow cylinder of barium titanate. The latter can be fairly remote from the orifice, since the only frequency distortion is that due to sound attenuation in the tube. This principle was first announced as long ago as 1933 by West,* who adapted the condenser microphone for the purpose, but it does not appear to have been exploited commercially.

Mr. D. E. L. Shorter: In the television service, the overriding requirements of the picture make it difficult to maintain the highest standards of sound quality. For example, an artist may be presented in a long shot with a scenic setting which does not allow a microphone to be placed at the normal distance without being seen: thus there is a need for microphones having highly directional characteristics substantially independent of frequency. For this reason some attention has been given during the past decade to the design of microphones responding to the second-order pressure-gradient in the sound wave; instruments of this kind are extremely costly to produce because they require two accurately matched elements, but they are at last becoming commercally available. An alternative approach is to avoid long-range working by concealing a microphone in the artist's clothing; some form of frequency-response correction is then necessary to compensate for the unnatural conditions of use.

With reference to the analogous circuit of Fig. 15, the diaphragm impedance, Z_1, can be represented as a load, and the rest of the circuit as a source having an impedance equal to Z_2 and Z_3 in parallel and an open-circuit 'voltage' which is a function of p_1, θ, Z_2 and Z_3. In this way, the fact, not always appreciated, that the directional characteristic is independent of Z_1 becomes evident by inspection.

Mr. R. B. Archbold: Although condenser microphones have been regarded in the past as the only satisfactory tool for laboratory precision measurements, nowadays, however, we see many more moving-coil microphones, and indeed use them very frequently for laboratory measurements. Can the author tell us something about the stability of the moving-coil microphone relative to the condenser microphone, and whether or not we can reasonably expect to use modern moving-coil microphones as laboratory standards?

Would the author comment on the frequency response of probe microphones, in particular on the response shown in Fig. 6, as it may give a false impression of present standards in this country? This represents the creditable performance of a probe microphone which can be turned out in bulk by the manufacturer. For laboratory work, probe microphones have been produced whose responses are substantially flat up to something like 10 kc/s.

Mr. H. D. Harwood: Can the author tell us something about the problems and tolerances involved in the manufacture of the different types of microphones he mentions and also indicate the degree of uniformity which may be expected in sensitivity and frequency response?

In Section 3.3 it is suggested that the microphone ribbon should be made as wide as possible to achieve high efficiency. In fact, there is an optimum width which is not critical and which depends to some extent on the criterion chosen, the decrease in efficiency of the magnetic circuit with width being counterbalanced by gains in other directions.

In Section 5 the author attributes the reduction in wind noise effected by a large windshield to the displacement of the principal components of the eddy noise to the sub-audible frequency range. An alternative explanation suggests itself, however, when it is remembered that eddies can be regarded as dipole or possibly quadrupole sound sources, producing a spectrum of which the low-frequency components decrease very rapidly with distance. Thus, whilst eddy noise generated at the microphone diaphragm contains a relatively high proportion of audible low-frequency components, that generated at the distant surface of a large windshield arrives at the diaphragm with those low-frequency components much attenuated.

Somewhat unexpectedly, a very high degree of windshielding can sometimes lead to difficulties. To speak in a high wind involves a considerable effort which is reflected in a change in speech quality. In the presence of some wind noise, this effect is accepted as normal in the circumstances; if, however, the wind noise is completely removed by a windshield, the quality may be critized as unnatural.

THE AUTHOR'S REPLY TO THE ABOVE DISCUSSION

Mr. M. L. Gayford (*in reply*): I am indebted to Mr. Clark for his interesting comments on directional microphones and their application to stereophonic techniques. In general the use of microphones in either the 'spaced microphone' or 'crossed microphone' stereophonic systems demands close matching both as regards frequency response and polar response. In reply to his question about the rating of microphones, I agree that the usual sensitivity rating giving the output in decibels relative to 1 volt per dyne/cm^2 fails to give information on the impedance characteristics, although it does give the open-circuit voltage generated. The impedance of most microphones varies considerably with frequency and any attempt to terminate the microphone in a 'nominal' resistance may result in a loss of efficiency and modification of the frequency response. The design of an optimum input transformer for the microphone amplifier requires the source impedance over the frequency range to be known. The microphone then works into a relatively high impedance and the maximum signal/noise ratio is achieved at the first valve grid. I feel that the practice of rating microphones in decibels referred to 1 mW with reference to a nominal impedance value can give misleading results owing to the difficulty of specifying the impedance. I prefer to state the open-circuit voltage generated by the microphone and to give the lowest permissible impedance into which the microphone can be worked without any appreciable modification of its response.

Mr. Clark rightly points out that microphones may be submitted to unexpectedly high sound levels at high frequencies. It is unfortunately very difficult to measure directly the distortion produced by high-quality microphones, as a distortion-free sound field is not easily produced. Constant-velocity transducers have distortion characteristics which are known to be a function of displacement and hence distortion normally falls rapidly as the frequency is increased. It is assumed that distortion is independent of frequency in constant-amplitude devices such as condenser microphones. Some measurements have been made which appear to confirm this.

* West, W.: 'A Device for Measuring Sound Pressures in Free Air', *Post Office Electrical Engineers' Journal*, 1933, 26, p. 260.

Mr. Robinson raises the question of the signal/noise ratio. The availability of better magnet materials and improved manufacturing techniques have tended to increase the efficiency of microphones, but, at the same time, there is a demand for further reductions in size in order to give less obtrusive microphones and more accurate high-frequency polar characteristics.

A probe microphone using a terminated pipe with an annular piezo-electric element as a section of the pipe is an attractive idea. It is possible that the device would be rather insensitive for use with small-bore pipes ($\frac{1}{4}$ in diameter or less).

Regarding Mr. Shorter's comments, it has so far proved very difficult to design highly directional microphones with a constant polar response at all frequencies.

In reply to Mr. Archbold, it appears that, given ordinary care in use and storage, moving-coil microphones make satisfacto laboratory standards. Condenser microphones may ha somewhat lower temperature coefficients, as it is possible construct them entirely of materials having nearly identic coefficients of expansion.

In reply to Mr. Harwood, space unfortunately do not permit a discussion on the manufacturing problems ar tolerances of various types of microphone. Condenser micr phones inherently require closer mechanical clearances tha ribbon microphones, but given equal care in manufacture bo types can show a high degree of uniformity of the order ± 1 dB in absolute sensitivity and response. Moving-co microphones have a more complex structure and hence it somewhat more difficult to control their characteristics.

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AN ELECTRON-TRAJECTORY TRACER FOR USE WITH THE RESISTANCE NETWORK ANALOGUE

By M. E. HAINE, D.Sc., Member, and J. VINE, M.Sc.

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SUMMARY

The paper describes an instrument for direct analogue computation of electron trajectories, a resistance network providing the necessary field data. Constructional details are given, and results for two typical electrostatic lenses are shown and compared with results obtained experimentally. Methods for improving accuracy and speed of operation are outlined.

LIST OF MATHEMATICAL SYMBOLS

V = Electrostatic potential.
V_0 = Potential at start of trajectory.
V_{min} = Minimum potential along trajectory.
V_N = Network electrode voltage.
r, θ, z = Cylindrical polar co-ordinates.
x, y, z = Cartesian co-ordinates.
u_r, u_z = Initial velocity components.
u_r', u_z' = Final velocity components.
u_{z0} = Velocity at start of trajectory.
u_{max} = Maximum velocity.
f_r, f_z = Acceleration components.
$\delta r, \delta z$ = Displacement components.
l_r, l_z = Network mesh dimensions.
$\delta V_r, \delta V_z$ = Network mesh voltages.
E_r, E_z = Electric field components.
e = Charge on electron.
m = Mass of electron.
t, t_{min} = Time.
$\left.\begin{matrix} V_u, V_{ur}, V_{uz} \\ V_u', V_{ur}', V_{uz}' \end{matrix}\right\}$ = Voltage analogues of velocity.
V_f, V_{fr}, V_{fz} = Voltage analogues of acceleration.
$\left.\begin{matrix} V_s, V_{sr}, V_{sz} \\ \bar{V}_{sr}, \bar{V}_{sz} \end{matrix}\right\}$ = Voltage analogues of displacement.
τ, τ_{min}, τ_0 = Mechanical analogues of time.
$\left.\begin{matrix} x_r, x_z \\ \bar{x}_r, \bar{x}_z \end{matrix}\right\}$ = Mechanical analogues of displacement.
m_r, m_z = Mesh voltage amplification factors.
$\left.\begin{matrix} a_r, a_z \\ b_r, b_z \\ \alpha \\ n_r, n_z \end{matrix}\right\}$ = Proportionality constants.
R_A, R_B = Resistances.
q^* = Focal point co-ordinate.
d = Principal surface separation.
r_0 = Value of r at start of trajectory.
f = Focal length.
u, v = Object and image distances.
S, T, D = Unipotential lens geometry parameters.

Dr. Haine was formerly and Mr. Vine is at the Research Laboratory of Associated Electrical Industries, Ltd. Dr. Haine is now at the Research Laboratory of Siemens Edison Swan, Ltd.

(1) INTRODUCTION

Many and varied types of apparatus have been developed for electron-trajectory tracing, and a good bibliography of some of them is given in the review paper by Liebmann.[1] An interesting recent example is the fully-automatic tracer described by Pizer, Yates and Sander,[2] which employs a special-purpose digital computer for integrating the equations of motion of the electron, the necessary field information being obtained from an electrolytic tank.

The advantages of accuracy and ease of operation that the resistance network has over the electrolytic tank, for the solution of electrostatic field problems, make it desirable that it should be employed in conjunction with some type of computer for plotting electron trajectories. The discontinuous nature of the resistance network introduces difficulties into the development of a fully-automatic device based upon it, but the analogue computer described in the paper promises to provide a means of semi-automatic plotting in which the duties of the operator will be reduced to the movement of a probe across the resistance network to feed the required field data into the computer.

(2) RESISTANCE NETWORK

The resistance network affords one of the most accurate and convenient methods for the determination of electrostatic fields with one degree of symmetry. The detailed theory of the network analogue has been well described[3,4] and only a brief outline will be given here.

In the plane type of problem normally described in terms of rectangular co-ordinates x, y, z, the potential V is independent of one co-ordinate, say z, and consequently the problem is solved if the potential distribution is determined over any single plane perpendicular to the z-axis. In the cylindrically symmetrical problem using cylindrical polar co-ordinates r, θ, z, the potential is independent of θ and it is sufficient to determine the distribution in any meridian plane, θ = constant.

As applied to either of these two types of problem, the resistance network provides a representation of the plane in which the potential distribution is required, divided into a large number of small rectangular meshes whose sides are parallel to the two co-ordinate axes concerned (Fig. 1). The sides of the rectangles are the resistors, and each nodal or mesh point, from which four resistors radiate, corresponds to a point in the plane under consideration.

To solve Laplace's equation

$$\nabla^2 V = 0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

for a simple problem of the potential distribution in charge-free space surrounding a system of electrodes, it is only necessary to represent the electrodes on the network by connecting up appropriate mesh points, and to apply voltages to these proportional to the actual electrode voltages. The potentials then arising at all the other network points are, to a good degree of accuracy,

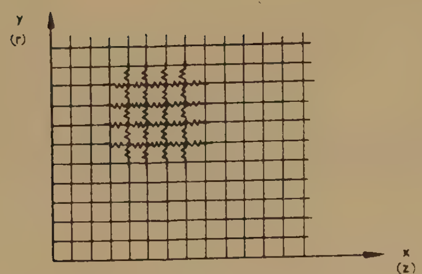

Fig. 1.—Portion of resistance network.

in the same proportion to those at corresponding points in the system under investigation, and can be measured at as many points as desired.

As an example, Fig. 12, to which detailed reference will be made later, shows how a symmetrical three-cylinder lens would be represented on an r, z network. For simplicity, the network is shown as a mesh of straight lines, and the mesh points to be connected together are shown joined by a heavy line. Owing to the rotational symmetry about the z-axis, each cylinder becomes simply a straight line and its representation on the network is very simple. Where the electrode cross-sections are curved, it becomes more difficult to fit them on the network. When the region of interest is fairly remote from the electrodes, it is usually sufficient to approximate to the curve by connecting together the mesh points which lie closest to it, but if potentials close to the electrodes are required, a more exact representation may be necessary and the boundary resistors must be adjusted.[3,4,5]

It should be mentioned here that a single network does not serve for the solution of both x, y and r, z problems. Whereas in an x, y network the value of a resistance is simply proportional to the length of the mesh side it represents, in an r, z network the relationship is different for the r and z directions and is also dependent upon the r co-ordinate.

In the more difficult type of problem where a space-charge distribution exists in the region around the electrodes, Poisson's equation,

$$\nabla^2 V = -4\pi\rho \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

must be solved, and the right-hand side is represented by feeding currents into the network at mesh points lying within the space-charge region. Owing to the interdependence of the potential V and the space-charge density ρ, an iterative process must be carried out, and except in certain trivial cases, the problem becomes much more complex.[4,6] Ray tracing will normally be involved in attempts to solve this type of problem, and the trajectory plotter is applicable to this in the same way as to the Laplace type.

A remaining point of note is that Liebmann has described a resistance network suitable for the solution of the class of magnetic-field problems in which the field is set up by currents flowing in coils concentric with the z-axis.[7] With slight modifications, which it is hoped to describe in a later paper, the trajectory plotter could be used to trace electron paths through such systems.

(3) BASIC THEORY OF THE TRAJECTORY PLOTTER

Although the plotter is potentially applicable to the more complicated space-charge and magnetic-field problems, for simplicity the paper is confined to consideration of ray tracing in the simple Laplace type of problem. The discussion will be carried out in terms of a cylindrically symmetrical problem using co-ordinates r and z, but it is equally applicable to a plane problem, and throughout one could read y for r and x for
The subscripts r and z will be used to distinguish between simila
physical quantities related to these two directions.

The computational method employed in the trajectory plotte
involves calculation of the electron path across the individua
small rectangles into which the network divides the field regio
the exit conditions of one rectangle providing the entry cond
tions of the next. Consider one such rectangle ABCD in Fig.

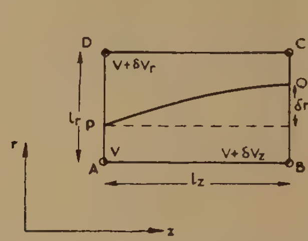

Fig. 2.—Enlarged drawing of a single network mesh.

where AB is parallel to the z-axis and AD is parallel to th
r-axis. Suppose the position of P, the entry point of the electro
to the rectangle, and the velocity components u_r, u_z of the electro
at this point, are known. Assume that the acceleration of th
electron is constant whilst it is in this mesh and that the accelera
tion components f_r, f_z are simply proportional to the voltages δV
δV_z which exist across the sides AD and AB; i.e. assume tha
the electric-field components E_r, E_z are given by

$$\left.\begin{aligned} E_r &= -\frac{\delta V_r}{l_r} \\ E_z &= -\frac{\delta V_z}{l_z} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad (3$$

where l_r, l_z are the lengths of the sides of the rectangle. The
Newton's second law gives

$$\left.\begin{aligned} f_r &= -\frac{e}{m}\frac{\delta V_r}{l_r} \\ f_z &= -\frac{e}{m}\frac{\delta V_z}{l_z} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad (4$$

These assumptions having been made without at present con
sidering the inaccuracies introduced by them, it becomes quit
simple to calculate the displacements δr, δz of the electron, a
any time t after entering the mesh, from the elementary equation
of dynamics. Thus we can write

$$\left.\begin{aligned} \delta r &= u_r t + \tfrac{1}{2} f_r t^2 \\ \delta z &= u_z t + \tfrac{1}{2} f_z t^2 \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5$$

Since in most lens problems the electron will be travelling mainl
in the z-direction, with small deviations in the r-direction, w
shall generally wish to calculate the value of δr shown in Fig.
after the electron has travelled right across the mesh in th
z-direction. To do this, put $\delta z = l_z$ in the second of th
equations (5) and solve the resulting quadratic equation for
This value of t is then substituted in the equation for δr and th
position of the exit point Q is thus determined.

Before repeating this for the next mesh, it is necessary to fin
the components, u_r', u_z', of the exit velocity. This is done b
inserting the value of t obtained above in the equations

$$\left.\begin{aligned} u_r' &= u_r + f_r t \\ u_z' &= u_z + f_z t \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad (6$$

u_r' and u_z then serve as the entry velocity components for the succeeding mesh.

Clearly, the fundamental factor affecting the accuracy of this method will be the size of the network meshes in relation to that of the electrode model. The larger the model can be made, in terms of network meshes, the greater becomes the number of steps in the computation; this means, in effect, that greater detail in field variation is taken into account, and, incidentally, the truncation errors in the potential solution itself become smaller. This question will be considered in the discussion of results in Section 10.

(4) DESCRIPTION OF THE COMPUTING CIRCUIT

The basic circuit of the analogue computing device is shown in Fig. 3. A voltage V_f is applied across a resistance R_0 in series with a potentiometer R_1, the potentiometer having a total resistance equal to that of R_0. To the output of R_1 is added a voltage V_u, obtained from a voltage source and potentiometer R_3, and the sum of these two voltages is applied across a second potentiometer R_2 whose output V_s can be measured on a bridge, formed by the voltage source V_b and potentiometer R_4, with the aid of the null indicator G. Potentiometers R_1 and R_2 are mechanically coupled so that the positions of their wiper contacts correspond, to a high degree of accuracy. We shall call τ the fractional setting of those potentiometers, i.e. τ is the ratio of the resistance tapped off by the wiper to the total resistance of the potentiometer.

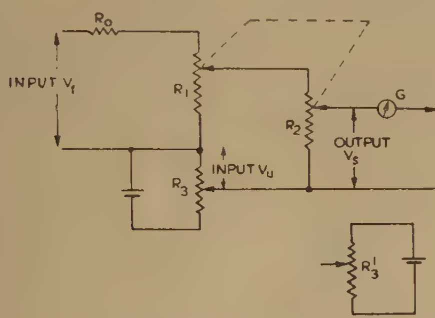

Fig. 3.—Basic computing circuit.

Since $R_0 = R_1$ it can be seen that, provided R_2 is of such a high resistance that the current flowing in it can be ignored, the voltage appearing across R_2 will be $V_u + \frac{1}{2}V_f\tau$ and thus the output V_s of R_2 will be given by

$$V_s = V_u\tau + \tfrac{1}{2}V_f\tau^2 \quad . \quad . \quad . \quad . \quad . \quad (7)$$

Comparing this with eqn. (5) it is seen that, if τ is made proportional to time t and V_u and V_f are proportional to the appropriate velocity and acceleration components, the output voltage V_s will be proportional to distance travelled in either the r- or z-directions.

To put this circuit to practical use in solving eqn. (5) we deal with the two components of displacement in turn. First, in the z-direction we wish to fix δz equal to l_z, the length of one network mesh, and to determine the time t necessary to cover this distance with given values of velocity u_z and acceleration f_z. It is part of the calibration procedure to determine the output voltage $\bar{V}_{sz}$ corresponding to a displacement l_z in the z-direction, and the output bridge is set for this value by adjustment of potentiometer R_4. (Calibration and setting-up is discussed in the Appendix and need not be considered here.) A voltage V_{fz} proportional to f_z is obtained from the network as described in Section 3, and a voltage V_{uz} proportional to u_z is fed in from potentiometer R_3. If the control of time potentiometers R_1 and R_2 is adjusted until the indicator G reads zero, the balance

$$\bar{V}_{sz} = V_{uz}\tau + \tfrac{1}{2}V_{fz}\tau^2 \quad . \quad . \quad . \quad . \quad (8)$$

holds and the value of τ then corresponds to the required value of t.

Having in effect solved the second of equations (5) for t, the next step is to insert the value of t in the first of these equations. Voltages V_{fr} and V_{ur} are therefore fed into the computing circuit, the setting τ being unchanged, and the output V_{sr}, given by

$$V_{sr} = V_{ur}\tau + \tfrac{1}{2}V_{fr}\tau^2 \quad . \quad . \quad . \quad . \quad . \quad (9)$$

is measured on the output bridge by adjustment of R_4, which is calibrated. The voltage V_{sr} is proportional to δr and can be converted into distance by means of the calibration figures.

The same value of τ must next be used for the derivation of voltages V_{ur}', V_{uz}', corresponding to the exit-velocity components of the electron. Fig. 4 shows the circuit rearrangement for this purpose. The resistor R_0 is short-circuited, the null indicator is transferred from the wiper of R_2 to that of R_1, and an auxiliary velocity potentiometer R_3', with its own voltage source, is connected as shown. R_3' is adjusted until G indicates zero, when the voltage output V_u' of R_3' is given by

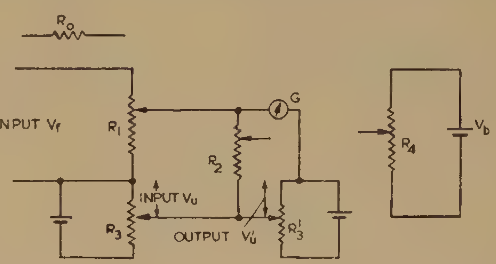

Fig. 4.—Rearrangement of circuit for velocity setting.

$$V_u' = V_u + V_f\tau \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Comparison of eqn. (10) with eqn. (6) shows V_u' to be the voltage required to represent the initial velocity in the next step of the calculation. R_3' is therefore retained at this setting and used in place of R_3 in the computation over the succeeding network mesh. Thus two pairs of velocity potentiometers are required, one pair providing the r- and z-velocity components for the current step in the plot, and the other pair standing by to receive the components for the following step.

An outline of the basic ideas having been given, the next Sections will deal with various aspects of the plotter in greater detail.

(5) SELECTION AND AMPLIFICATION OF NETWORK VOLTAGES

The basic idea propounded in Section 3, that the voltages across sides AD and AB of the rectangle ABCD shown in Fig. 2 should be taken as being proportional to the field components within this rectangle, although convenient for simplicity of explanation, is not the most satisfactory way of using the network voltages. More correctly, in a step-by-step computation, for the calculation of an element PQ of the trajectory one requires a good estimate of the field components at the mid-point of this element, but owing to the discontinuous nature of the resistance network, voltages representing these components will not generally be directly available. Without introducing extra complications into the method, the best one can do is to use voltages

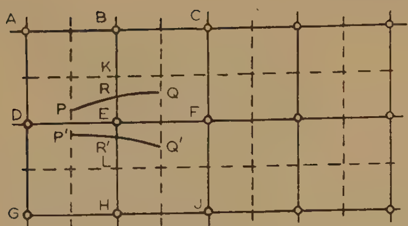

Fig. 5.—The use of network voltages for particular trajectory elements.

representing the components at the nearest available points to the mid-point of PQ.

Fig. 5 illustrates a possible system in which the elementary steps in the trajectory are displaced from the network meshes by a half mesh in the z-direction. Here A, B, C . . . J are nodal points of the network and the dotted lines join the centre points of the meshes. For the calculation of the element PQ of the trajectory, instead of the field components at R, one uses δV_{DF}, the voltage between points D and F, for the component E_z (this voltage would really be applicable to the field at point E), and voltage δV_{EB} for component E_r (really the value at K). For a path element such as P′Q′ the same voltage δV_{DF} would represent E_z, but E_r would now be given by δV_{HE}.

In a case where the path element cuts the line DF it is usually a simple matter to estimate beforehand whether it will do so to the left or to the right of the point E. This decides which of the two voltages δV_{EB} and δV_{HE} is the more appropriate one to represent E_r.

It can be seen that, with this method, the voltages used give an approximation to the field components at points displaced slightly, in the r-direction, from the mid-point of the trajectory element.

The probe used to pick off the network voltages makes contact with five nodal points, e.g. B, D, E, F and H in Fig. 5. The potentials from B, E, H are fed into a switch on the control panel which allows selection of either δV_{EB} or δV_{HE} for E_r, according to the considerations described above. δV_{DF} is fed in directly for use in the z computation.

A d.c. supply of 4–20 volts is used to energize the resistance network, and the resulting mesh voltages are quite small, varying from a maximum of around half a volt down to microvolts. The voltages therefore require amplification before being used in the computing circuit, and initially this was accomplished by means of the simple multiplying device shown in Fig. 6. In this,

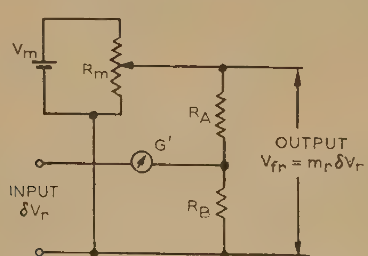

Fig. 6.—Manual multiplying circuit.

potentiometer R_m is adjusted until the null indicator G′ reads zero, when the output V_{fr}, which is being fed into the computing circuit, is m_r times the input δV_r from the network, m_r being determined by the ratio of resistances R_A and R_B:

$$m_r = \frac{R_A + R_B}{R_B} \qquad . \quad . \quad . \quad . \quad . \quad (11)$$

A similar circuit employed for the z-direction usually requires different multiplying factor, m_z.

The manual balancing operation necessary in this device ha now been eliminated by means of an electronic instrumen developed by R. Thorn from his previously described electroni null-indicator.[8] The circuit is basically that of the manua device, but the output is obtained by amplification of the out-o balance signal from the galvanometer. Fig. 7 is a schematic c

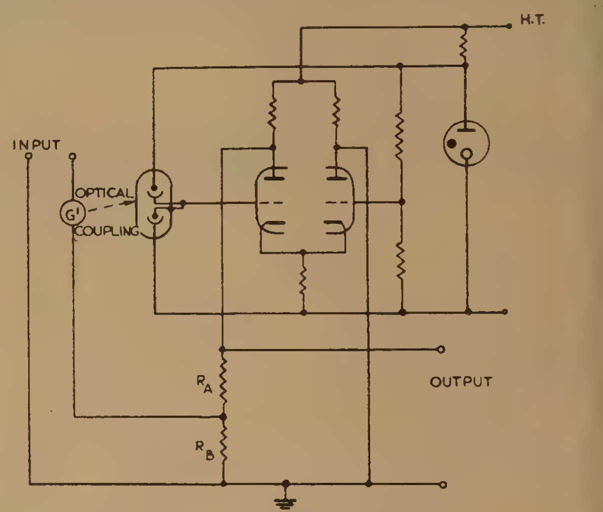

Fig. 7.—Galvanometer-amplifier (schematic).

the new arrangement. The light reflected from the galvanc meter falls upon a split-cathode photo-electric cell whose tw sections are connected in series. The output voltage from th junction of the two sections is fed to one grid of a balanced d. amplifier whose other input is a fixed voltage taken from voltage divider across the h.t. supply. Output is taken via balanced cathode-follower, omitted for clarity from Fig. 7 t the resistance chain R_A, R_B in such a sense as to oppose th original galvanometer deflection. The mechanical alignment c the system is adjusted so that at zero input the two sections c the photocell are equally illuminated, and the voltage divide providing the reference level is set to give zero output unde these conditions.

A very high loop gain, estimated at 2×10^{11} volts output pe ampere input, enables the residual off-balance voltage to b reduced to negligible proportions, although this voltage mus of course, be finite in order to provide an input signal to th amplifier. The overall amplification of the system is still dete mined by resistances R_A and R_B and is within $0{\cdot}1\%$ of th value given by eqn. (11), for output voltages approaching th maximum of 20 volts that the instrument is designed to giv and for values of m_r, m_z between 2 and 1 000. The instrumer has proved reasonably insensitive to vibration when a fir mounting is provided for the galvanometer unit, and zero dri is less than 10 μV, referred to the input, over a period of sever days.

To set the desired values of the amplification factors m_r, m resistances R_A and R_B are varied by means of a set of wir wound resistors and a system of short-circuiting links, enablir R_B to vary from 1 ohm to 1 111 ohms in step of 1 ohm, whil R_A is such as to keep the total, $R_A + R_B$, equal to 10 kilohm The resistors are not accurately made to their nominal value and it is necessary to measure m after it has been set to rough the required figure. This is done by connecting a standa decade potentiometer as shown in Fig. 8. The output of t galvanometer amplifier is balanced against the total volta

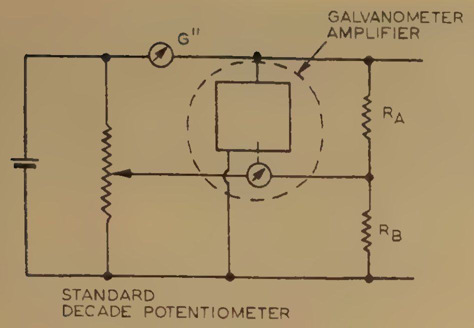

Fig. 8.—Circuit for measurement of m_r or m_z.

across the standard potentiometer, the reading of which gives accurately the ratio $R_B/(R_A + R_B)$.

A single galvanometer amplifier is used to amplify the *r*- and *z*-network voltages in turn, and as different amplification factors are normally required for the two, a resistance chain such as R_A, R_B is included for each direction, the amplifier being switched to the appropriate one as its input voltages are switched.

(6) SWITCHING CIRCUIT AND VOLTAGE SUPPLIES

Fundamentally the circuit comprises two of the basic computing circuits shown in Fig. 3, one for each of the *z*- and *r*-directions, but by operating these in turn, the duplication of certain equipment, such as power supplies and the photocell amplifier, is avoided. It is also convenient to use a single time-potentiometer section (components R_0, R_1, R_2) to avoid the complication involved in coupling two such systems mechanically. A control switch is incorporated to permit the interchange of shared components, between the *z*- and *r*-circuits, in a single operation.

A second control switch is used to facilitate the rearrangement of the circuits from the displacement-measuring form (Fig. 3) to the velocity-resetting form (Fig. 4), whilst a third switch enables the velocity potentiometers R_3 and R_3' to be interchanged rapidly at the completion of each step in the plot.

The power unit for the computing circuits comprises three similar stabilized voltage supplies mounted on one chassis. Two of these supply the velocity potentiometers and one the output bridge. Each supply is variable between 50 and 100 volts and is designed to feed a fixed load of about 1 kilohm. Stability against mains fluctuations is of the order of 0·1%. Generally, a lower voltage level exists in the *r*-circuit than in the *z* and, to make best use of the potentiometers, lower supply voltages are desirable. Voltage-dropping resistors are normally connected in series with the *r*-velocity potentiometers and output bridge, so that no adjustment of the supply voltage is necessary on switching from *z* to *r*. Additional parallel resistors are also inserted to keep the overall load on the supplies near the optimum value.

An additional stabilized supply is necessary to energize the resistance network. The power requirement is normally small as the impedance between the network electrodes is usually 50–100 ohms and voltages below 20 volts are used. A small stabilized supply using transistors is employed.

(7) TIME POTENTIOMETERS AND OUTPUT BRIDGES

It was decided to aim at an accuracy of 0·1% in the computing circuits so that the following requirements had to be met by the time potentiometers:

(*a*) Resolution greater than 1 000.
(*b*) Linearity 0·1%.
(*c*) Accurate coupling so that the wipers of R_1 and R_2 track together within 0·1%.
(*d*) R_2 to have approximately 1 000 times the overall resistance of R_1.

To satisfy this specification, a Kelvin–Varley slide system is used, employing switched resistors in preference to continuous toroidal potentiometers, to make possible the accurate coupling. Each potentiometer comprises a coarse and fine switch control, the switches each having 50 positions giving a resolution of 2 500. The linearity and accurate coupling are assured by the use of high-stability wire-wound resistors of adequate power rating and a tolerance of 0·1%. Requirement (*d*) was met by making R_1 of total resistance 1·25 kilohms and R_2 1·25 megohms; this keeps the output impedance of the system down to a reasonable figure.

For the output bridge potentiometers a similar resolution and linearity are required as for the time potentiometers, and the same system is used. The total resistance of both *r*- and *z*-potentiometers is 2·5 kilohms, and each is supplied with a reversing switch enabling both positive and negative output voltages to be measured.

Calibrated dials fitted to the time potentiometers and output bridges allow readings to be taken as four-figure decimals, 1·0000 being the maximum on each dial. The resolution of 2 500 means that calibration markings go in steps of four in the fourth decimal place.

(8) VELOCITY POTENTIOMETERS AND NULL DETECTOR

As it was not considered necessary to take readings of velocity components, the linearity of these potentiometers is not important and the only requirement is that the resolution shall be sufficient to obtain an accurate balance. A form of Kelvin–Varley slide is used, employing a 2-gang wire-wound potentiometer as the coarse control, with a fine control connected between the two wipers. The arrangement is shown in Fig. 9. A

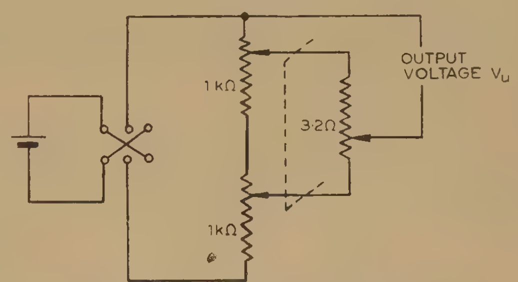

Fig. 9.—Detail of a velocity potentiometer.

reversing switch for each velocity unit allows representation of positive or negative velocities.

This system is not altogether satisfactory, as it has been found difficult to ensure sufficiently high stability in the coarse-control wiper contact. When the wiper is set very close to a turn of wire, but not firmly on it, the electrical contact between the two may be intermittent, causing the output voltage of the potentiometer to fluctuate. It is intended to eliminate this trouble by fitting switched coarse controls, as used in the other potentiometers.

When in use in the circuit of Fig. 3, the velocity potentiometer is subject to the loading effect of the second time potentiometer R_2, and to allow for this a resistance of similar magnitude is connected across the output of the reserve potentiometer R_3' (Fig. 4) so that the conditions under which the new velocity voltage is set on this potentiometer are similar to those it will experience when in use. Early tests carried out without this dummy load showed that the velocity-setting error was appreciable, and being always in the same direction it is cumulative.

A valve voltmeter serves as the null indicator G, employed in all the balancing operations. This has a sensitivity of approximately 30 mV per scale division near the null point, and an out-of-balance as small as 5 mV can therefore be detected. A protection device causes the sensitivity to decrease rapidly for high input voltages, so that large off-balance signals up to 100 volts can be observed without damage to the meter, and no range switch is necessary.

(9) GENERAL PLOTTING PROCEDURE

To commence a trajectory plot through an electrode system of given geometry, the initial data required are the starting co-ordinates and velocity components of the electron. Once the electrode model has been set up on the resistance network, the calibration procedure described in the Appendix is carried out to determine the constants relating output bridge readings to displacement values in the two co-ordinate directions. The z-output bridge is then set at a reading corresponding to a distance of one mesh. The pair of velocity potentiometers to be used in the first step is selected and set to voltages corresponding to the initial velocity components.

Each step in the plot involves four balancing operations, one for the setting of time t, one for measurement of δr and two for the velocity resetting. The successive δr values are summed so that the exact value of the r-co-ordinates is known at the start of each step, and whenever this value passes through a multiple of half a mesh-width the probe connections are moved in the r-direction.

(10) RESULTS

Preliminary tests were carried out by plotting electron paths in a region of constant field strength. For this particular problem, errors due to finite mesh size are zero, so that it is a useful means of testing the accuracy of the computing circuits. Paths corresponding very closely to the expected parabolas were obtained, the discrepancy between the experimentally determined displacement and that calculated being in most cases as low as 0·1%, even after 20 to 30 steps.

It was thus possible to consider more complicated problems with non-uniform fields, in the knowledge that any inaccuracy in the results would be due to the finite size of the network mesh and the consequent crudity of the approximation to the values of the field components acting on the electron.

A good source of results for comparison purposes is Liebmann's paper on unipotential electrostatic lenses.[9] The ideas necessary for understanding these results are illustrated in Fig. 10,

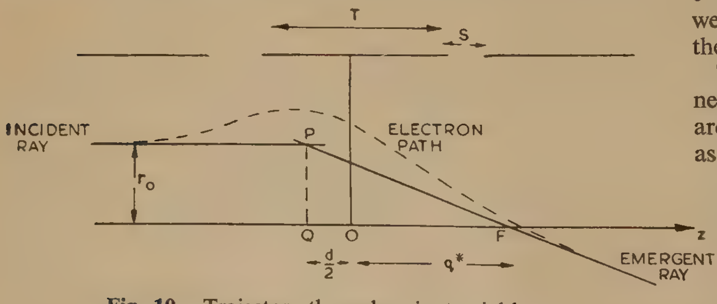

Fig. 10.—Trajectory through unipotential lens.

where a typical electron path through such a lens is shown by the broken line. The emergent direction of the electron is produced back, giving the focal point F where it cuts the axis, and the position of a point P on the principal surface where it cuts the incident ray. PQ is perpendicular to the axis. We will use Liebmann's notation in writing q^* for OF and $d/2$ for OQ. His results, obtained experimentally, are presented in the form of graph for each lens, showing the variation of $d/2$ and q^* with r_0 the initial off-axis distance of an electron moving parallel to th axis. The paraxial focal length f is given by $q^* + d/2$ evaluate for a ray very close to the axis. Liebmann's curves for tw lenses are shown in Fig. 11, where the points marked by circle show results obtained with the trajectory plotter.

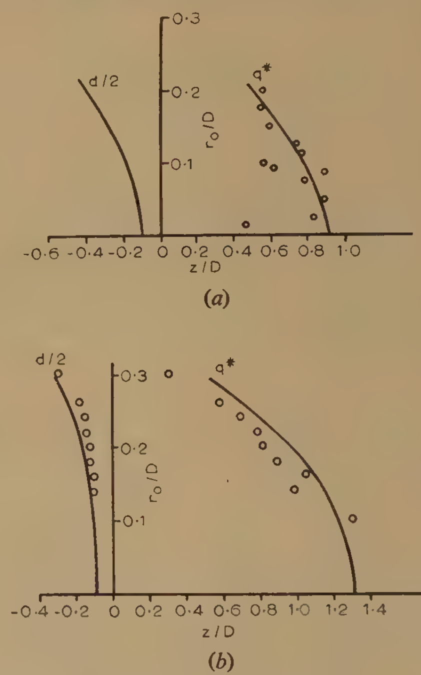

Fig. 11.—Results for lenses.

(a) Lens 5. T/D = 0·667, S/D = 0·5, D = 18 meshes.
(b) Lens 3. T/D = 0·6, S/D = 0·133, D = 30 meshes.

The resistance network used in these tests is that describe by Liebmann in his 1955 paper.[4] It has a square fine mes with 60 × 20 unit steps in the z- and r-directions respectivel extended to 180 × 50 in coarser steps. The two lenses, number 5 and 3 in Liebmann's numbering, were set up as shown i Figs. 12(a) and 12(b) respectively. The representation of lens gives ratios S/D and T/D in exact agreement with Liebmann figures, whilst for lens 3 there may be a slight discrepancy i that S/D = 0·133 for the network model against the figure o 0·13 stated by Liebmann. The outer cylinders of both lense were extended to the edges of the network, no dimensions fo them being given by Liebmann.

To take account of as much of the lens field as possible, it necessary that the plots should start and finish in regions whic are practically field-free. This presented no difficulty with lens as, with the scale used, the edges of the fine mesh, $z = \pm 3$ were found to be at about 0·995 of anode potential, so th these were suitable points at which to start and end th plots. A much larger scale was decided upon for lens and the model was off-set to the right of the network shown in Fig. 12(b). The field at the left-hand edge of th fine mesh was then negligible, and plots were carried o from left to right as far as lens centre and from right to le thereafter, so as to start and end at this edge.

The seriousness of the inaccuracy introduced by the finite mes size is shown by the results in Fig. 11(a), which have very litt relationship to a smooth curve. By repeating several plots, was confirmed that this scatter was not due to computation errors. The reproducibility was found to be good, discrepanci

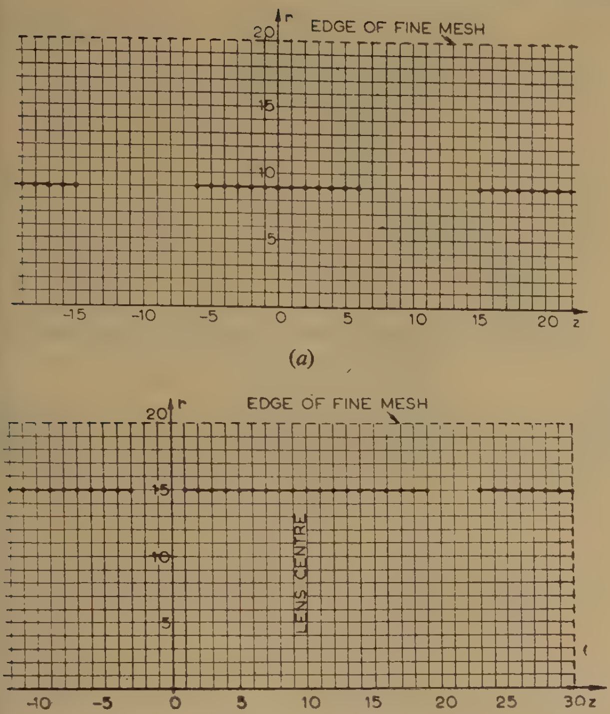

Fig. 12.—Lenses as set up on the network.
(a) Lens 5.
(b) Lens 3.

being roughly of the order of the diameter of the circles on the graph. Location of the principal surface demands fairly accurate determination of the emergent ray slopes, and so no results for this are shown in Fig. 11(*a*).

The results for lens 3 in Fig. 11(*b*) show much less scatter, and the points for large values of r_0 approximate quite well to a curve of the required shape. The fact that this curve does not agree with Liebmann's may be due to various factors. The lens used was perhaps a little stronger owing to the possible discrepancy in cylinder separation ($S/D = 0{\cdot}133$ instead of $0{\cdot}13$); although the effect of this would probably be small, it would be in the right direction to contribute to the observed discrepancy. In addition, Liebmann's results were obtained not by the use of incident rays parallel to the axis but by calculation from results obtained with an object at a finite distance, using the ideal lens formula $-1/u + 1/v = 1/f$.

Agreement over the position and curvature of the principal surface of this lens is satisfactory, and indicates the greater accuracy of these results.

(11) CONCLUSIONS

The results obtained are sufficient to show the limitations of this method of trajectory plotting, and to indicate where improvements must be made to achieve greater accuracy. The way in which the mesh size introduces inaccuracies has been explained in Section 5, where it was seen that, in the computation of any path element, the field component values used are not generally those at the centre of the element, as desired, but those at some point, or points, having a slightly different r co-ordinate. The maximum possible discrepancy in the r co-ordinate is half a mesh, $l_r/2$, and it is therefore this dimension which is mainly responsible for the errors. An obvious improvement would be to have a rectangular network mesh with l_r smaller than l_z, instead of the square mesh of the network used for these experiments. This finer subdivision of the network mesh in the r-direction would allow the variation of field strength with r to be more accurately taken into account in the computation, without the necessity of an impossibly large network model. It would be mainly required in the region close to the axis, for the mesh-size error decreases further out, as can be seen from the following consideration.

For axially symmetric fields, if terms in r^2, etc., are ignored,

$$E_r \propto r \qquad (12)$$

$$E_z = \text{constant} \qquad (13)$$

for a fixed value of z (see, e.g., Reference 10). This means first that the mesh size mainly affects the accuracy of E_r and is relatively unimportant as far as E_z is concerned. Secondly, it follows from eqn. (12) that

$$\frac{dE_r}{E_r} = \frac{dr}{r} \qquad (14)$$

so that if dr is the maximum possible discrepancy between the radial co-ordinates of the point at which E_r is required and the point at which it is determined, this is constant and equal to half a mesh, and the percentage error dE_r/E_r is therefore proportional to $1/r$. This is borne out by the results shown in Fig. 11, where the amount of scatter is noticeably greater for the smaller values of r_0. It was for this reason that no plots were made for lens 3 for values of $r_0 < 0{\cdot}1$, or three meshes, where the inaccuracy is becoming marked.

A network designed specifically for this method of trajectory plotting could, therefore, have a graded mesh, l_r being small at the axis, say equal to $l_z/10$, and increasing smoothly with r. This would enable accurate results to be obtained with conveniently small electrode models, even for paraxial rays, which cannot be satisfactorily traced with the network at present available.

An alternative procedure to building a special resistance network is to use some system of interpolation to obtain a more accurate value for the field components. Since, as has been stated, E_r is approximately linear with r, a fairly accurate system could be arranged involving measurement of E_r at two points with different r co-ordinates, the value of E_r to be used in the computation being determined from the two measured values by linear interpolation, taking into account the exact position of the electron between the two points. Such a system is being considered and it is hoped to report on this development in the future.

No experiments have been carried out for the two-dimensional x, y type of problem, but it is likely that the effect of large mesh size will be much the same. Accuracy is likely to depend a great deal on the nature of the problem. In the r, z case, mesh subdivision in the r-direction becomes necessary because, generally, the path is roughly parallel to the z-axis and crosses comparatively few meshes in the r-direction. Because the electron crosses the complete mesh in the z-direction, errors in E_z tend to cancel out over each individual step, which does not apply to the r-computation. If, in an x, y problem, the paths were roughly at 45° to the axes, the cancelling of errors over each step might apply in both component directions so that a more reliable result could be expected. Since the choice of axes is arbitrary in an x, y problem, it would be possible to arrange the electrodes so that the trajectories passed diagonally across the network, thus giving maximum accuracy.

As well as the modifications required for increased accuracy, it is intended to introduce automation of the balancing operations, by means of servo mechanisms, in order to make plotting more rapid. The function of the operator will then be to move the probe to the appropriate network point and initiate the balancing sequence at each step. A system for the addition of

successive δr voltages, without the need for taking readings, can also be incorporated and made to give an indication when the probe connections must be moved in the r-direction. With the addition of these refinements, the plotter could become a convenient and quick means of obtaining accurate trajectory plots through electrostatic fields. In its present manually-operated form, the time taken for a single plot over about 50 steps is of the order of two hours, but with automatic balancing and complete separation of the r and z circuits, each with its own power supplies, etc., so that two balancing operations can be made simultaneously, this might be reduced to as little as half an hour.

(12) ACKNOWLEDGMENTS

The instrument described is based on an idea which was jointly conceived by the late Dr. G. Liebmann and one of the present authors (M. E. H.). The basic system was worked out before Dr. Liebmann's death.

Thanks are due to Mr. R. T. Taylor for assistance with much of the detailed design, to Mr. R. Thorn for the development of most of the electronic equipment used, and to Dr. T. E. Allibone, Director of the Research Laboratory, Associated Electrical Industries, for permission to publish the paper.

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(14) APPENDIX: CALIBRATION AND SETTING-UP

Subscripts r and z will be used to distinguish the symbols for similar quantities in the two directions.

Suppose the acceleration components f_r, f_z are represented in the computer by voltages V_{fr}, V_{fz} and velocity components u_r, u_z by voltages V_{ur}, V_{uz}. Let the displacement components δr, δz be represented by voltages V_{sr}, V_{sz}.

Express the acceleration and velocity proportionalities by the equations

$$V_{fr} = a_r f_r \quad V_{fz} = a_z f_z \qquad (15)$$

$$V_{ur} = b_r u_r \quad V_{uz} = b_z u_z \qquad (16)$$

The setting τ of the time potentiometers is proportional to time t.

Let
$$\tau = \alpha t \qquad (17)$$

Considering the z-direction first, we have eqn. (8):

$$V_{sz} = V_{uz}\tau + \tfrac{1}{2}V_{fz}\tau^2$$
$$= (b_z\alpha)u_z t + (a_z\alpha^2)\tfrac{1}{2}f_z t^2$$

so that
$$V_{sz} = a_z\alpha^2\delta z \qquad (18)$$

provided we arrange that

$$b_z = a_z\alpha \qquad (19)$$

The relationship between output voltage V_{sz} and displacement in the z-direction is determined by eqn. (18), provided we know a_z and α, but these constants cannot be conveniently measured and the method of obtaining them is now described.

Let V_N be the network supply voltage, V the voltage at some point on the network and, as before, δV_r, δV_z the voltages representative of the acceleration components at this point.

Let
$$u_{max} = \sqrt{\left(\frac{2eV_N}{m}\right)} \qquad (20)$$

be the velocity the electron would have if accelerated to full network potential,* and let

$$\tau_{min} = \alpha t_{min} \qquad (21)$$

where t_{min} is the time required to travel a mesh length l_z in the z-direction at this speed.

Then
$$l_z = u_{max}t_{min} \qquad (22)$$

To fix α we choose a convenient value for τ_{min}, bearing in mind that τ must never exceed unity at any point or it would be impossible to set on the time potentiometers. It is necessary, then, to estimate V_{min}, the lowest potential likely to be encountered by an electron in the field, and to fix $\tau_{min} < \sqrt{(V_{min}/V_N)}$ which ensures that, when the electron is travelling at its slowest, a value of τ somewhat less than unity is obtained. It is assumed here that the electron always travels in the positive z-direction and never reverses. Special consideration would be required for such a trajectory, such as dividing the path into sections and calibrating for each part separately.

Substituting in eqn. (22) for u_{max} from eqn. (20) and for t_{min} from eqn. (21), we have

$$l_z = \frac{\tau_{min}}{\alpha}\sqrt{\left(\frac{2eV_N}{m}\right)}$$

Therefore
$$\alpha = \frac{\tau_{min}}{l_z}\sqrt{\left(\frac{2eV_N}{m}\right)} \qquad (23)$$

so that, having chosen τ_{min}, α is now fixed.

To find a_z, note that eqn. (4) gives

$$f_z = \frac{e}{m}\frac{\delta V_z}{l_z}$$

and that
$$V_{fz} = m_z\delta V_z \qquad (24)$$

m_z being the amplification factor used in the z-circuit.

Thus
$$V_{fz} = m_z l_z \frac{m}{e} f_z$$

which compared with eqn. (15) shows that

$$a_z = m_z l_z \frac{m}{e} \qquad (25)$$

* We can speak as though V_N were the electrode voltage in the actual problem, since the voltage does not affect the trajectory, and we avoid the introduction of another proportionality constant.

Putting values of α and a_z given by eqns. (23) and (25) into eqn. (18), we have for the relationship between V_{sz} and δz

$$V_{sz} = m_z l_z \frac{m}{e}\frac{\tau_{min}^2}{l_z^2}\left(\frac{2eV_N}{m}\right)\delta z$$

$$= 2m_z\tau_{min}^2 V_N \frac{\delta z}{l_z} \quad . \quad . \quad . \quad . \quad . \quad . \quad (26)$$

In practice we do not wish to make absolute measurements of voltages. The output V_{sz} is obtained as a reading, say x_z, on the z-output bridge dial, and we wish to determine the value to be set on this dial, say $\bar{x}_z$, to correspond to the displacement $\delta z = l_z$. The voltage $\bar{V}_{sz}$ corresponding to the reading $\bar{x}_z$ is

$$\bar{V}_{sz} = \bar{x}_z V_{bz} \quad . \quad . \quad . \quad . \quad . \quad . \quad (27)$$

where V_{bz} is the total voltage across the bridge. Knowledge of the absolute value of this voltage or of V_N is unnecessary if V_N is measured directly on the z-output bridge, a reading, n_z say, being obtained:

$$V_N = n_z V_{bz} \quad . \quad . \quad . \quad . \quad . \quad . \quad (28)$$

We can now put eqns. (27), (28) and the condition $\delta z = l_z$ into eqn. (26) and obtain

$$\bar{x}_z = 2m_z n_z \tau_{min}^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (29)$$

in which τ_{min} is a chosen constant and m_z and n_z are found by measurement, so that $\bar{x}_z$ can be calculated.

Similar considerations apply to the r-direction, where, corresponding to eqns. (26), (27) and (28), we have

$$V_{sr} = 2m_r\tau_{min}^2 V_N \delta r \frac{l_r}{l_z^2} \quad . \quad . \quad . \quad . \quad . \quad (30)$$

$$\bar{V}_{sr} = \bar{x}_r V_{br} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (31)$$

$$V_N = n_r V_{br} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (32)$$

so that bridge readings x_r can be converted into displacements by the equation

$$\frac{\delta r}{l_r} = \frac{x_r}{2m_r n_r \tau_{min}^2}\left(\frac{l_z}{l_r}\right)^2 \quad . \quad . \quad . \quad . \quad . \quad (33)$$

while the reading corresponding to the distance l_r would be

$$\bar{x}_r = 2m_r n_r \tau_{min}^2 \left(\frac{l_r}{l_z}\right)^2 \quad . \quad . \quad . \quad . \quad . \quad (34)$$

In order to commence a trajectory plot one must calculate $\bar{x}_z$ from eqn. (29) and the constant relating δr to x_r in eqn. (33). Of the various constants involved, τ_{min} is obtained from the underestimate of V_{min}/V_N which is generally simple; n_z and n_r are found directly by measurement of V_N on both r and z output bridges, facilities for this being provided; l_z/l_r is a constant of the resistance network; and m_z and m_r may be determined by means of the circuit shown in Fig. 8, as described in Section 5.

These measurements having been made, it remains to set the initial velocity components. For the sake of example, assume the initial velocity to be parallel to the z-axis:

$$u_{z0} = \sqrt{\left(\frac{2eV_0}{m}\right)} \quad . \quad . \quad . \quad . \quad . \quad (35)$$

where V_0 is the potential at the starting-point. If τ_0 is the time-potentiometer setting corresponding to the time taken to travel the distance l_z at this speed, we have

$$\frac{\tau_0}{\tau_{min}} = \frac{u_{max}}{u_{z0}} = \sqrt{\left(\frac{V_N}{V_0}\right)}$$

so that

$$\tau_0 = \frac{\tau_{min}}{\sqrt{\left(\frac{V_0}{V_N}\right)}} \quad . \quad . \quad . \quad . \quad . \quad . \quad (36)$$

The ratio V_0/V_N is determined by a simple bridge measurement on the resistance network and τ_0 is calculated. With the setting $\bar{x}_z$ on the z-output bridge, zero acceleration input, and τ_0 set on the time potentiometers, the velocity potentiometer is adjusted to give a balance (see Fig. 3), and the required setting is then obtained.

A trial plot may be necessary to decide whether the values of any of the constants can be changed with advantage. Thus, if the r-output readings seem to be small, so that the resolution available on the output bridge is not realized, the value of m_r may be increased; if too small a range is utilized on the time potentiometers, the underestimate of V_{min}/V_N might be revised.

DISCUSSION BEFORE THE MEASUREMENT AND CONTROL SECTION, 7TH APRIL, 1959

Dr. K. F. Sander: The paper records an interesting development on the subject of relating a quantized system to a continuously variable one, and the authors are to be congratulated on the ingenious solution which they have found. Before discussing their method, I must say that in my opinion the statement concerning the greater accuracy of the resistance network over the electrolytic tank is unjustified. The form of tank described by Mr. Yates and myself gives accuracies in field measurement at least as great as those derived from resistance networks using 0·1% resistors.

Why do the authors use a d.c. supply for the network? The use of a.c. would make the processes of amplification and buffering much easier. The square-wave measuring technique used on our electrolytic tank set-up was derived from that described by Yates, Lucas and Johnston* in connection with strain-gauge resistance bridges. I would not recommend the use of sinusoidal a.c. for resistance bridges of this type. The point about using square-waves is that the pickup always affects the leading edge and not the centre at all. It appears to me eminently suitable for a resistance network.

* YATES, J. G., LUCAS, D. H., and JOHNSTON, D. L.: 'Multi-Channel Measurement of Physical Effects by Confluent Pulse Technique, with particular reference to the Analysis of Strain', *Proceedings I.E.E.*, Paper No. 1032 M, October, 1950 (98, Part II, p. 109).

I should like further information on the results recorded in Section 10. Can the authors describe the 'parabolic' tests in further detail? In particular, how were the parabolae oriented, and what does the 0·1% signify geometrically? It is essentially a one-dimensional form of check, and as such, the results may not be immediately applicable to fields varying in two dimensions. The statement later in Section 10, that the consistency of the lens results was indicated by the circles in Fig. 11, appears to mean about 3% of focal length, and may be an indication of random mounting errors in a truly two-dimensional field. We found, when using a tank and mechanical integrators, that results were much more critically dependent on integrator performance than we would be led to expect from static accuracy figures. Could something be said about the relative magnitudes of the acceleration and velocity contributions to the displacement in the individual steps?

Finally, in taking trajectories, the authors have evidently thought it proper to work at intervals much less than a mesh side. In fact, 0·2 and 0·3 are used. It appears likely that if a given trajectory crosses a fair number of radial meshes, systematic errors may tend to cancel. If it is confined to a small number, larger effects might be expected. Could the authors say, for example, how far the paraxial trajectories in Fig. 11(*a*) departed

from the axis, and whether the violent fluctuations in the focal points are associated with crossing into a new radial mesh interval?

Mr. P. Muff: Can the authors indicate how the accuracy and speed of their method compare with previous methods which one has to use with resistance networks, that is, taking field data, making a potential map of the system and hand plottin trajectories across it?

Mr. A. H. Doveton: Have the authors considered correction for relativistic effects and the application of the plotter t problems where this would be of importance?

THE AUTHORS' REPLY TO THE ABOVE DISCUSSION

Messrs. M. E. Haine and **J. Vine** (*in reply*): Regarding the points raised by Dr. Sander, an analysis of the accuracy of solutions obtained by resistance network has been made by Liebmann (Reference 4 in paper). Three kinds of error arise, namely measurement error, errors due to resistor tolerance and truncation error. Liebmann's work and our experience indicate that, for a network built from 0·1% tolerance resistors with mesh size reduced so as to render truncation error negligible, potential values would be obtained, with residual errors arising from the other two sources, of between one part in 10^4 and one in 10^5. It seems likely that measurement error would limit the electrolytic tank to a lower accuracy than this.

The question of the ultimate accuracy obtainable is, however, academic, since one is usually concerned with the amount of effort required to obtain results of a given accuracy. On this basis, we can be more definite regarding the network's superiority, as the arguments advanced by Liebmann* in its favour have not, to our knowledge, been refuted.

Both a.c. and d.c. supplies have been used in resistance-network applications and experience has shown the latter to be preferable. The use of square-wave excitation would probably be satisfactory, but it is unlikely to have a balance of advantage over direct current, for which the stabilization of supplies is more straightforward. Amplification by means of the galvanometer amplifier has proved quite satisfactory.

In the parabolic tests queried by Dr. Sander, an electron travelling in the x-direction was considered to enter a uniform field parallel to the y-axis, producing a parabolic track with vertex at the point of entry. Analytical treatment, for the field and velocity values chosen in one case, indicated that the electron should pass through the point $x = 16$, $y = 10 \cdot 667$ (meshes), referred to the vertex as origin. Results of two runs carried out with the plotter were, at $x = 16$, $y = 10 \cdot 647$ and $10 \cdot 663$ (0·2% and 0·04% accuracy respectively). Results of similar accuracy were obtained for other field and velocity values. The purpose of these tests was to determine the limit of accuracy obtainable with the equipment, and they were performed with great care and as rapidly as possible so as to minimize power-supply variation.

In assessing results there are two aspects to be considered; first the consistency of different results obtained with the plotter for the same trajectory, and secondly the agreement of the plotter results with the true solution, if this is known. The consistency of results is dependent only upon the accuracy with which voltages can be set and measured and maintained stable (the measurement accuracy), whereas agreement with the true solution is affected, in general, by discrepancies between the true field values acting on the electron and the values used in the computation, according to the considerations of Sections 5 and 11. The latter effect will be termed the 'position' error, since it is the result of determining the field in the wrong place.

In a uniform field, the position error is zero. Since the field is the same everywhere, no inaccuracy is introduced by determining it in the wrong place and it would be more appropriate to call this a zero-dimensional test. The agreement of the parabolic results with the true result as well as their consistency is therefore dependent only on the measurement accuracy. The indicate the ability to set and measure voltages to within abou 0·1%, and to maintain them stable to this degree for sho periods, say $\frac{1}{2}$ h.

With similar care and speed, the consistency of results for an field distribution (variation in one or two dimensions) would b expected to be comparable with this figure and in some cases i was. However, less care was taken in obtaining the result shown in Fig. 11, since the main interest at that stage was in th effect of the position error. These results indicate the consistenc to be expected from the device in normal operation.

With regard to the relative magnitudes of r- and z-components the best indication is given by comparing the amplificatio factors m_r, m_z, used for the two circuits, eqn. (11). Usually m_z is 5–10 times m_r; for example, $m_z \simeq 5$, $m_r \simeq 50$. I extreme cases, the velocity components may become of the sam order, as in the outermost trajectory of Fig. 11(b) which has a average slope of about unity on its convergent section.

The initial divergent effect in the decelerating section of unipotential lens is quite small and the maximum excursion o the trajectories from the axis is approximately equal to the initia value, r_0, plus 20%. It should be emphasized that the results i Fig. 11(a) were taken close together and near to the axis on small-scale model, in order to illustrate the effect of the positio error most clearly. Thus, the two lowest points are for trajec tories which do not depart by even one mesh from the axis, an only for the top three points do the trajectories exceed $r =$ meshes. From Fig. 11(b), it is seen that the worst effects ar avoided if the paths cover about six meshes in the radia direction.

For the trajectories starting close to the axis the position erro is largest, as shown by eqn. (14), and also it is more systemati because these paths move so slowly in the r-direction. Conside for example, a trajectory starting at $r = 0 \cdot 8$ mesh which reache the strongest part of the radial field with, say, $r = 0 \cdot 95$ (approxi mately 1 mesh, but slightly less). Because of the linear rise o E_r with r, the field value used in the computation, measured a $r = 0 \cdot 5$ mesh, is approximately half the correct value. The sma slope of such a trajectory means that the electron passes righ through the most important field region (extending over just few meshes) with a practically unchanged r-co-ordinate, an consequently there is a systematic build-up of error and the tot radial impulse computed will be roughly half what it shoul be. On the other hand, an electron starting with $r = 0 \cdot 85$ mes may pass through the strong-field region at a distance of slightl more than 1 mesh from the axis, say $r = 1 \cdot 05$ mesh. The radi field for this would be measured at $r = 1 \cdot 5$ mesh, a value of 1 times the true value, giving in this case a systematic error at th opposite extreme. The values for the focus position obtaine from two such plots would differ greatly. Effects of this typ are responsible for the wide scatter of results in the paraxi region.

The errors for extra-axial plots are much more random, sin these trajectories have a comparatively large slope and m cross several radial intervals in traversing the most importa field region.

For lens investigations one is most interested in the close paraxial paths, in the region $r/D < 0 \cdot 1$. The problem chose

* LIEBMANN, G.: 'Electrical Analogues', *British Journal of Applied Physics*, 1953, 4, p. 193.

here for the purpose of illustration is therefore of the type to which the trajectory tracer is least suited. When not required to operate in the paraxial region, the device provides reasonable accuracy and has given satisfactory service in investigation of space-charge-limited electron guns.* Also, a simple interpolation system, enabling the plotter to give improved accuracy in the paraxial region, has been demonstrated, and is described elsewhere.†

In reply to Mr. Muff, the trajectory-tracer method compares favourably with other plotting methods with which we have had experience. For hand-plotting methods, the preliminary recording of field data is time-consuming, whether it be in the form of equipotentials for graphical methods, or in numerical form for computational methods. In place of this there is the calibration procedure to be carried out, but this will require significant time only when one is faced with an unfamiliar problem.

The actual trajectory tracing by graphical methods (e.g. the well known 'parabola method') can be rapid enough to make the total time taken, including the mapping of equipotentials, comparable with that taken by the hand-operated trajectory tracer. The accuracy, however, is poor, and also the methods can break down if faced with equipotentials of awkward though not unusual shape.

Hand-computational methods of equivalent accuracy to the trajectory tracer generally take considerably longer. The only exception to this might be for the case of paraxial rays in which a solution of the paraxial equation by the method, for example, described by Coslett (see p. 37 of Reference 10 in the paper) can be quite quick, and in which, to obtain reliable results from the trajectory plotter, it would be necessary to use the interpolation system mentioned above. Even in this case it is doubtful if the overall time, including the recording of field data, could be reduced to that required with the plotter.

In check computations, trajectories have been traced using the actual trajectory plotter method, as described in Section 3, with a hand calculating machine. The time taken for a single track was about four hours, plus up to an hour for recording the field data (the latter time being dependent upon the radial traverse of the trajectory). This is more than twice the time taken using the trajectory tracer and greater effort and concentration were demanded from the operator.

Finally it should be mentioned that methods have now been developed for the application of an electronic digital computer to the ray-tracing problem.‡ These are able to use to the full the field-data accuracy obtainable from existing resistor networks, and virtually reduce the time taken to that required for the conversion of the field data from analogue to digital form. Given an efficient and accurate means for performing the latter operation, these methods promise to be at least as accurate as, and more convenient than, any existing methods.

In reply to Mr. Doveton, correction for the relativistic increase in mass of the electron at high accelerating potentials has been considered, and a possible means of accomplishing it can be described briefly.

Making allowance for the relativistic effect, eqns. (4) are replaced by

$$\left.\begin{aligned} f_r &= -\frac{e}{m_t}\frac{\delta V_r}{l_r} \\ f_z &= -\frac{e}{m_l}\frac{\delta V_z}{l_z} \end{aligned}\right\} \quad \ldots \ldots \quad \text{(A)}$$

* ARCHARD, G. D.: 'Trajectory Plotting in Electron Guns', *Proceedings of the Physical Society* (to be published).
† VINE, J., and TAYLOR, R. T.: 'An Improvement to the Electron Trajectory Tracer' (to be published).
‡ VINE, J.: 'Application of a Combination of Analogue and Digital Computers to Electron Trajectory Tracing' (to be published in *The Computer Journal*).

where the transverse mass

$$m_t = \frac{m}{(1 - u_z^2/c^2)^{1/2}}$$

and the longitudinal mass

$$m_l = \frac{m}{(1 - u_z^2/c^2)^{3/2}}$$

c being the velocity of light and m the rest mass of the electron.*

Strictly, these equations apply only when the electron is travelling, instantaneously, parallel to the z-axis, and so consideration is limited to problems where this condition is at least approximately true along the entire path, e.g. paraxial conditions in an electron-microscope lens. The assumption is made that the electron mass is constant across each mesh step in the trajectory, just as the acceleration was assumed constant in writing eqn. (5).

Eqns. (A) can be taken into account by supplying voltages V_{fz}, V_{fr}, to the trajectory plotter, not proportional to δV_z, δV_r as in eqn. (24), but proportional to $\delta V_z\,(1 - u_z^2/c^2)^{3/2}$ and $\delta V_r\,(1 - u_z^2/c^2)^{1/2}$ respectively. To a good degree of approximation, for accelerating voltages of about 50 kV (velocity less than $\frac{1}{2}c$), the binomial factors can be replaced by $1 - \frac{3}{2}u_z^2/c^2$ and $1 - \frac{1}{2}u_z^2/c^2$ respectively, and simulated by means of potential-divider networks of the form shown in Fig. A. Here,

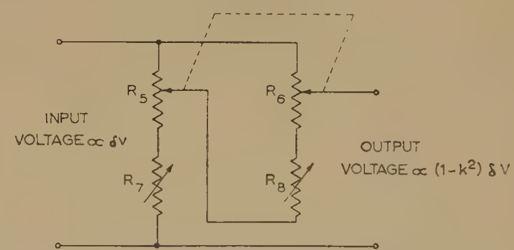

Fig. A.—Divider network for relativistic correction.

potentiometers R_5 and R_6 are coupled together, and the second divider, R_6, R_8, is of sufficiently high resistance for it to have negligible loading effect on R_5, R_7. The voltage across the second divider is a fraction, k, of the input voltage, and the voltage between the wiper of R_6 and the upper input terminal is a fraction, k^2, of the input voltage. Consequently, the output voltage, taken between the wiper of R_6 and the lower terminal, is proportional to $(1 - k^2)\delta V$. For the z-computation, k is made equal to $\sqrt{\frac{3}{2}}u/c$, and for the r-computation, to $\sqrt{\frac{1}{2}}u/c$, the circuits being inserted between the network probe and the trajectory-plotter input terminals. In practice, it might be necessary to introduce an electronic buffer stage between the network and the correction circuit to eliminate undesirable loading effects.

If R_5 and R_6 are also coupled to the z-velocity potentiometer, the correction adjustments would be automatically made as the velocity potentiometer is set at each step. R_7 and R_8 are necessary because the maximum traverse of the velocity potentiometer will not, in general, correspond to the velocity of light, and they would be pre-set at the start of a trajectory so that

$$\frac{R_7}{R_5 + R_7} = \frac{R_8}{R_6 + R_8}$$

$$= \sqrt{\frac{1}{2}}\,\frac{u_{max}}{c} \text{ for the } r\text{-circuit}$$

and

$$= \sqrt{\frac{3}{2}}\,\frac{u_{max}}{c} \text{ for the } z\text{-circuit.}$$

* See, for example, McCREA, W. H.: 'Relativity Physics' (Methuen, 1935), p. 52.

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HIGH-POWER TELEVISION TRANSMITTERS FOR BANDS IV AND V

By T. S. ROBSON and T. M. J. JASKOLSKI, Dipl.-Ing., Associate Members.

(*The paper was first received* 24*th April*, 1958, *in revised form* 14*th January, and in final form* 25*th May*, 1959.)

SUMMARY

To increase the number of channels available for television broadcasting, it is now becoming increasingly necessary in several countries to utilize the ultra-high-frequency bands (Band IV, 470–585 Mc/s, and Band V, 610–960 Mc/s). The paper considers some of the factors concerning the relatively restricted coverage on ultra-high frequencies and the power requirements of transmitters to offset this disadvantage. This leads to a discussion on the choice of valve types for use in u.h.f. television transmitters. The choice of a high-gain klystron output amplifier for a 10 kW transmitter enables a relatively low-power modulated amplifier to be used in the penultimate stage. The paper describes the fully engineered transmitter installed experimentally at the B.B.C. television transmitting station at Crystal Palace for propagation and television reception tests in the u.h.f. band covering both the 405-line and 625-line television systems.

(1) INTRODUCTION

Frequency Bands IV and V have been allocated internationally for television broadcasting, but they have not yet been exploited to any extent in Europe. They cover the frequencies 470–585 Mc/s and 610–960 Mc/s, a total of 465 Mc/s. Using a 5 Mc/s channel width in accordance with the existing British 405-line television standards, 93 u.h.f. channels would be available. An increase in channel width in the United Kingdom is practicable only in Bands IV and V because the channel spacing in these bands has not yet been internationally allocated for the European zone. The present C.C.I.R. Western European channel width is 7 Mc/s, and its increase to 8 Mc/s in Bands IV and V is probable.

A fundamental disadvantage of the u.h.f. channels for television broadcasting is that their propagation characteristics are basically inferior to those obtainable with the v.h.f. channels. Of necessity, use is now being made of ultra-high frequencies for television in the United States, but, so far, we in the United Kingdom have little practical experience of its potentialities, and such experience is essential to the proper planning of a u.h.f. network.

It is likely that in the not too distant future at least part of the u.h.f. bands will be used for television broadcasting in this country. There is great interest in u.h.f. television on the continent of Europe, especially in Germany, where a few low-power transmitters (about 1 kW output) are already in experimental operation.* The paper describes a vision and sound transmitter which has been built to obtain experience in designing for the relatively high powers which will be needed at ultra-high frequencies and to provide a source of transmission for a study of propagation, coverage and reception problems at these frequencies.

These transmitters were installed at the B.B.C. television transmitting station at Crystal Palace and started radiating regularly for test purposes on the 11th November, 1957. Th transmissions up to 27th March, 1958, were made on the Britis 405-line standard at 654·25 Mc/s (vision) and 650·75 Mc/ (amplitude-modulated sound). Transmission on the C.C.I.R Western European 625-line standard at 654·25 Mc/s (vision) an frequency-modulated sound at 659·75 Mc/s started on 5th May 1958, and were continued until the end of August, 1958.

(2) TRANSMITTER POWER REQUIRED

The general pattern of regional planning in the Unite Kingdom is such that on the v.h.f. channels of Bands I and III few high-power transmitting stations are used to give coverag to a large proportion of the population, and the remaining gap in coverage are then filled in by low-power transmitting stations At ultra-high frequencies a considerably greater effective radiate power (e.r.p.) is required (see Table 1), and it is interesting t note that in the United States the Federal Communication Commission have raised the limit for a u.h.f. station t 5 MW e.r.p.[1]

Table 1

Tentative Comparison of Performance of Band V (600 Mc/s 1 MW e.r.p. and Band I (50 Mc/s) 100 kW e.r.p.

	Band I 50 Mc/s	Band V 600 Mc/s	Relative gain or loss for Band V
			dB
Transmitting station e.r.p.	100 kW	1 MW	+10
Receiver aerial wavelength factor	—	—	−22
Receiver aerial gain ..	3 dB	8–12 dB	+ 7 (average)
Receiver noise factor ..	6–8 dB	12–16 dB	− 7 (average)
Receiver aerial mismatching and feeder loss	2 dB	5 dB	− 3
		Total	−15 dB (average)

(*a*) *Receiver aerial wavelength factor*.—The signal voltage across the output terminal of a half-wavelength receiving aerial is proportional to the wavelength, resulting in factor of −22 dB (approx.) for 600 Mc/s, compared with 50 Mc/s.
(*b*) *Receiver aerial gain*.—About 8 dB is common in the u.h.f. bands, while larg arrays give up to 12 dB.
(*c*) *Receiver noise factor*.—A good commercial receiver has a noise figure of abou 14 dB at u.h.f., but 8 dB can be achieved with a low-noise r.f. stage. The Table give more conservative figures between 12 and 16 dB.
(*d*) Field strengths required for, and factors affecting, reception in various tele vision bands are dealt with in detail elsewhere.[2]

It is desirable that the service areas of u.h.f. stations shoul be comparable with those of the equivalent high-power Band stations, but this is unlikely to be attained as 1–2 MW appear to be the practical limit to station e.r.p. at present and tha already accounts for higher transmitting aerial gain at th higher operating frequencies. At ultra-high frequencies th maximum net transmitting aerial power gain, allowing for losses that can be effectively used is about 20, and for 1 MW e.r.p. 50 kW transmitter would therefore be required. Assuming transmitter of this power can be made, it should be possible t operate two of them in parallel to give 2 MW e.r.p.

When our transmitter development started (1953), valves wit

* At the time of publishing the paper two 10 kW u.h.f. transmitters are also operated in Germany: one by the Südwestfunk[19, 21] and one by the Fernemeldetechnisches Zentralamt.[20, 22]

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Mr. Jaskolski is, and Mr. Robson was formerly, with E.M.I. Electronics, Ltd. Mr. Robson is now with the Independent Television Authority.

a maximum c.w. output of 50 kW were not available, and the highest-power valve commercially available at that time, giving 10 kW, was used. Using an aerial with a gain of 20 and allowing for losses, an e.r.p. of approximately 150 kW became practicable.*

The associated amplitude-modulated sound transmitter is required to give the same peak power at 100% modulation, corresponding to a carrier power of 2·5 kW. A frequency-modulated sound transmitter operates at 1/5 of the vision peak power (C.C.I.R. Western European standards), i.e. 2 kW output.

(3) CHOICE OF VALVE FOR THE POWER AMPLIFIER

The proposed operating frequency was above the maximum usable frequency of any high-power triode or tetrode available at the time. Transit-time effects and r.f. losses tend to make these types of valve unsuitable for use in the upper range of the u.h.f. band, but these difficulties have largely been defeated in certain specialized high-power u.h.f. ceramic tetrodes now becoming available.[3, 4, 20, 22] While ultra-high frequencies undoubtedly complicate the design and manufacture of valves of conventional type, such frequencies permit the use of valves, such as klystrons, that rely on transit time for their operation. The types of transmitting valve that can be used[4] are listed below.†

(3.1) U.H.F. Tetrodes

Some operating data of three u.h.f. tetrodes (available in 1953), types 6181, 6448 and 6806, are given in Table 2.

It will be noted that either type 6448 or 6806 would give the power specified for the transmitter described. The construction of these two valves is interesting in that the cathodes, grids and anodes are concentric, with the anode on the inside. This construction makes it possible to use them in a simple circuit with an earthed cathode and a driven grid. A greater power gain can then be obtained than is possible with the earthed-grid cathode-driven arrangement, which is usual at higher frequencies. Type 6448 in this circuit can give a power gain of about 12.

The type 6181 valve is of too low power for the output stage of the transmitter concerned, but it would be useful as a driver valve for either of the other two. Type 6448 will give a peak output of 12 kW throughout the frequency range required, and is therefore suitable. Type 6806 was not available when the design of this transmitter was started, and it is noted that, although it gives appreciably greater power output at the lower frequencies of the u.h.f. band, its performance at the upper end of the frequency range is not significantly better than that of the 6448.

Thus, of these valves, type 6448 was a possible choice. The advantages of using a valve of this type are:

(*a*) For these frequencies the efficiency is reasonably high, being approximately 50%.
(*b*) The efficiency is maintained over all brightness levels of the picture.
(*c*) High-level modulation can be employed by modulating the output amplifier.
(*d*) With the use of a single type of valve, the circuits can be made to tune over the whole u.h.f. band.

The disadvantages are:

(*a*) A very elaborate mechanical construction of the circuit arrangement is necessary, and great care has to be taken to avoid feedback from the output to the input circuits.
(*b*) The circuit must contain several decoupling capacitors, which are a possible source of unreliability.
(*c*) As the power gain is only about 12 it is necessary to use valve type 6181 as a driver stage. This valve requires almost as elaborate a circuit as does the output valve.
(*d*) The close spacing of the control grid and cathode in a high-power valve detracts from its reliability. A flash-over that damages the grid is likely to cause a grid-cathode short-circuit.
(*e*) At the higher frequencies, these valves will provide powers of the order of 10 kW, but much larger output powers than this will later be required.
(*f*) Water cooling is required.
(*g*) For the wide bandwidths required, the use of double-tuned circuits is necessary, which complicates the mechanical design and increases the cost of the amplifiers.

Table 2

DATA OF U.H.F. TETRODES TYPES 6181, 6448 AND 6806 (MAXIMUM RATINGS)

Valve type	Anode direct voltage	Anode dissipation	Bandwidth at −3 dB	Power output into load					
				Peak synchronizing pulse level (negative modulation)			Peak white level (positive modulation)		
				At 500 Mc/s	At 750 Mc/s	At 900 Mc/s	At 500 Mc/s	At 750 Mc/s	At 900 Mc/s
	kV	kW	Mc/s	kW	kW	kW	kW	kW	kW
6181	2	2	8	—	—	1·2	—	—	0·8
6448	7	26	7	15	—	12	9·8	—	7·8
6806	9	36	7	28	17	—	18	11	—

(3.1.1) Tetrodes in Bridge Circuit.[22]

Recently, two or four tetrodes in a bridge circuit have been used to double or quadruple the available power output from one valve. By this means an amplifier with an output power of the order of 10 kW can be built using relatively small valves.

(3.2) Klystrons[1, 5, 6, 7, 8, 9, 19, 21]

At the start of our development work, two high-power klystrons were commercially available with an output of about 10 kW. They were the 3K50,000L and the GL–6237–6242. The first type has ceramic windows across the cavity gaps so that the cavities can be in part external to the vacuum. This leads to a flexibility of design, as the cavity and coupling arrangements are under the control of the transmitter designer, and enables the whole of Bands IV and V to be covered with only three valve types.

In the second type of klystron, the cavities are wholly contained within the vacuum system of the valve. Control of operating frequency is obtained by deflecting the walls of the cavities, but this gives only a limited frequency range and six klystron types are necessary to cover Bands IV and V. This type has the

* For the Crystal Palace Band V installation, an allowance of 1·5 dB must be made for losses in the combining filter, feeder and waveguide. Thus, with a 10 kW transmitter and aerial gain of 20, this transmitting station has an actual e.r.p. of 143 kW and the expected disadvantage compared with a 100 kW Band I station is then 23·5 dB.

† For description and characteristic features of u.h.f. transmitting valves see also Reference 5, which gives a comprehensive bibliography of the subject.

advantage that there is no ceramic material in the r.f. fields in the cavities, which could be a source of trouble.

The advantages of a klystron are:

(a) The klystron is basically a very simple structure with no closely spaced electrodes. It fits into a circuit consisting only of a number of resonant cavities surrounded by focus coils. No decoupling capacitors are necessary. This leads to a greater reliability.
(b) The cavities are well spaced and there is small risk of feed-back between them. Neutralizing arrangements are therefore not required.
(c) The power gain of a three-cavity klystron tuned for maximum bandwidth is approximately 100. For 10 kW output the driver stage thus has to provide only about 100 watts. The use of the 1 kW penultimate stage needed to drive the u.h.f. output tetrode is thus avoided.
(d) The beam current drawn from the h.t. supply is constant, and thus contains no video-frequency components. It is therefore not necessary to provide constant-resistance smoothing filters in this supply.

The disadvantages of the klystrons are:

(a) Several different types are required to cover the whole frequency range.
(b) Klystrons being large and heavy are not easy to handle.
(c) Magnetic circuits are required to focus the beam.
(d) The average efficiency is lower than that of tetrodes or triodes.
(e) At present, water cooling is required above 2 kW output.

(3.3) Travelling-Wave Amplifiers

These valves are another possibility. Their efficiency is likely to be similar to that of the klystrons and the r.f. circuits should be simple. As high-power valves of this type were not commercially available, they were not considered for this transmitter.

(3.4) Magnetrons

High-power magnetron valves have been made. These are essentially oscillators and cannot readily be amplitude-modulated or used as linear amplifiers. They could possibly be used with an absorption modulation system, but this would be difficult to apply. It would also be necessary to employ an automatic frequency-control system.

(3.5) Final Choice of Valves

After considering the foregoing possibilities, the final decision was made in favour of klystrons on the grounds that they are more suitable for the whole u.h.f. range, need less complicated and cheaper circuits, and can be developed to give greater power output than is obtainable from tetrodes. By carefully considering the mechanical design of the circuit and by providing suitable handling facilities it was decided that the size and weight of the klystron would be no great drawback. The provision of a suitable magnetic circuit for focusing the beam did not prove either difficult or expensive, and setting up and operating the klystron proved to be no more difficult than for a conventional valve stage.

Of the two types of klystron available, it was decided to use the one with external cavities, as the design of these cavities is more flexible and it is possible to load them as required.

Fig. 1 shows the klystron used. It employs a bombarded-type cathode in which a tantalum block is bombarded by electrons from a heated tungsten filament maintained at 2·3 kV negative with respect to the cathode. The beam of electrons emitted from the cathode is first focused electrostatically and then further focusing is done magnetically as the beam travels along the tube, by a number of coils energized from a 100-volt d.c. source. Four of them are situated inside a magnetic cage. The coil nearest to the cathode, and the most critical one in setting-

Fig. 1.—Klystron type 3K50,000LF (overall length approximately 4 ft)
Eitel McCullough Inc.

up, is known as the prefocusing coil; it is located just befor
the point where the electron beam enters the magnetic cage
The last focus coil is between the output cavity and the collecto
The collector is near earth potential and the cathode is main
tained at approximately −17 kV. The water-cooled collector i
capable of dissipating 50 kW.

(3.6) Progress in Klystron Development

During the last five years considerable progress has bee
achieved in klystron development. The paper does not attemp
to discuss this matter in general as it is dealt with elsewhere.[5, 2]
However, it should be noted that, to obtain both higher gain an
larger bandwidth, the tendency is to increase the number o
resonant cavities; for example, 4-cavity types with integral an
with external cavities are now available. Klystrons with a
output above 10 kW and 6-cavity power klystrons have bee
developed, but those commercially available at present are, t
the best of our knowledge, outside the frequency range o
Bands IV and V.

It is interesting to note that the new 10 kW 4-cavity klystron
have an increased frequency coverage, as follows:

With built-in cavities:
Type VA-833A covers 685–985 Mc/s (part of Band V).
Type VA-833B covers 470–685 Mc/s (Band IV and part o
Band V).

With ceramic windows and external cavities:
Type 4KM50,000 LA covers 400–610 Mc/s (Band IV).
Type 4KM50,000 LQ covers 610–985 Mc/s (Band V).

The latter types are provided with a modulating anode which, if connected through a high resistance to earth (collector), considerably limits the arcing current in case of an internal flash-over. Moreover, the bombarded-type cathode has been replaced in these new klystrons by a sintered 'matrix-type' unipotential cathode which works at a relatively low temperature and avoids the complication of the bombarded-type cathode, as well as providing a longer life expectation.

(4) VISION TRANSMITTER

(4.1) Power Amplifier[1, 7, 8]

(4.1.1) General.

The klystron is mounted with the collector uppermost in a steel-sided frame. This frame, which also contains the resonant cavities and focus coils, is mounted on wheels and can be removed from the cubicle after disconnecting the water and supply circuits by quick-release couplings. A schematic of the klystron amplifier is shown in Fig. 2.

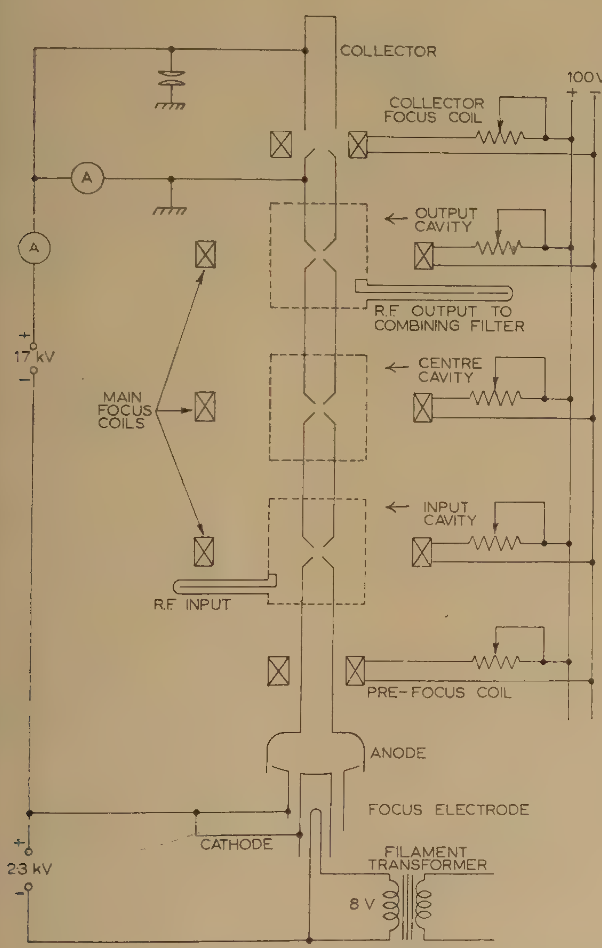

Fig. 2.—Klystron amplifier schematic.

(4.1.2) Magnetic Circuit.

As mentioned above, four or five focus coils (depending on the operating frequency) are used. Along the length of the klystron a flux density of the order of 100 gauss is required. Round the klystron, just above the cathode, is a small pre-focus coil providing a magnetic lens which forms the electrons into a parallel beam. Round the main body of the klystron are two or three large coils of approximately 24 in diameter, each providing about 4 000 AT. The coil diameter is large to allow room inside for the r.f. cavity boxes, and also to ensure that a fairly uniform field is maintained along the axis of the klystron. The final focus coil is close to the collector to ensure that the beam enters it correctly. Fig. 3 shows the graph of the magnetic flux

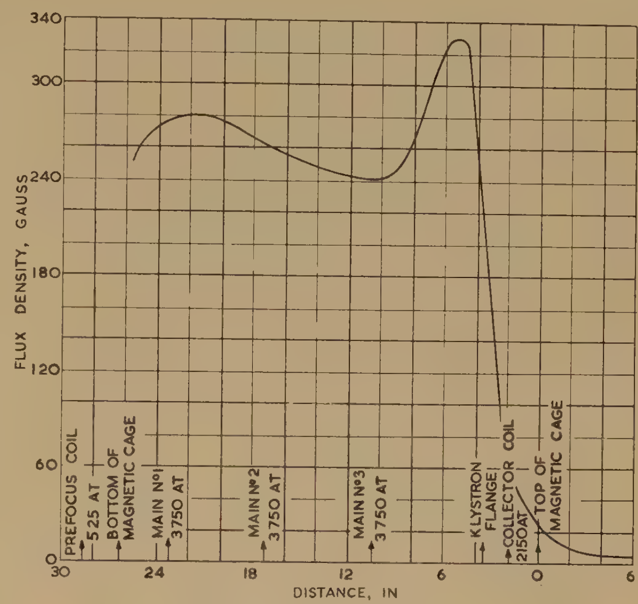

Fig. 3.—Magnetic flux density along klystron axis.

density along the klystron axis (klystron removed) for a typical setting of the focus coils (all coils energized and producing: 525 AT pre-focus; 3 750 AT each main coil; 2 150 AT collector coil).

The current through each focus coil is independently adjustable to give a minimum reading on the body-current meter which indicates that a minimum number of electrons are striking the walls of the klystron. When the body current is a minimum a beam transmission ratio, (collector current)/(beam current), of 95% can be obtained. Under normal conditions the klystron is operated with a slightly lower beam transmission ratio to obtain a greater r.f. power output. This is because a wider-diameter beam couples more closely to the r.f. fields in the cavities, but is closer to the walls of the gaps.

The focus coils are fed from a common 100 volt 12 amp d.c. supply employing metal rectifiers.

(4.1.3) Resonant Cavities.

The theory of re-entrant cavities leads to rather complicated formulae for their resonant wavelengths.[9, 10] As a crude approximation, however, one can calculate the resonance from the gap capacitance and the inductance, considering the cavity as a single-turn toroidal coil. The height of the cavity being determined by the height of the klystron's ceramic window, prototype cavities of reasonable proportions were first built and tested before finalizing the cavity design. Fig. 4 shows the frequency coverages of two such cavities, both having the same height of $4\frac{1}{2}$ in and separation of sliding tuning faces varying between 6 and 12 in. Curve (*a*) shows the resonant frequencies for $8\frac{1}{2}$ in distance of fixed walls, whilst (*b*) applies to a fixed-wall spacing of 15 in. Curve (*a*) very nearly corresponds to the frequency coverage of the cavities as finally used in the transmitter ($8\frac{1}{2} \times 12 \times 4\frac{1}{2}$ in high).

It is important that the tuning arrangement does not upset

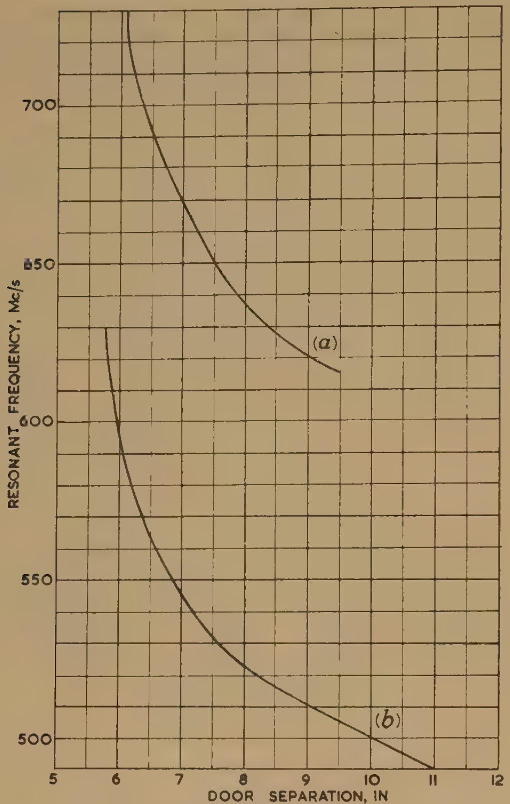

Fig. 4.—Frequency coverage of klystron cavities.
(*a*) Height $4\frac{1}{2}$ in, width $8\frac{1}{2}$ in.
(*b*) Height $4\frac{1}{2}$ in, width 15 in.

the symmetrical nature of the cavity, otherwise the alignment of the r.f. field in the klystron cavity gap may be affected. The end faces are fitted with spring fingers and can be moved in or out by lead-screws. The cavity boxes are split and slide together to make contact through spring fingers with the klystron cavity flanges.

Experiments were made with a circular cavity resonator 11 in diameter and $4\frac{1}{2}$ in high. It was tuned by means of fifteen 2×4 in vanes mounted round the circumference in such a way that they could be rotated as required.[11] A tuning range of 605–707 Mc/s was obtained. The advantage of this shape of cavity is that it takes up less space than the rectangular type. This is especially important at Band IV lower frequencies, where a cavity diameter of 17 in is necessary, and where rectangular cavities are difficult to fit into the space available within the diameter of the magnetic focus coils. Rectangular cavities were in fact used in this transmitter because they proved to be easier to manufacture.

Rotatable loops are provided in the cavities for coupling to external loads or sources. The coupling to the input cavity is adjusted to give low standing-wave ratio on the transmission line between the modulated amplifier and the klystron. As the output coupling is increased from zero, the output power rises to a maximum and then decreases. The normal working position of the loop is slightly overcoupled, i.e. rotated some 5° to 10° past the maximum output position. This ensures that a safe and stable operating condition is achieved without risk of excessive voltage being produced across the cavity gap. For 405-line working, adequate bandwidth can be obtained by stagger-tuning the centre cavity. For greater bandwidth it is necessary to employ damping loads coupled to the input cavity, to the centre cavity, or possibly to both. Stagger-tuning of the centre cavity and the introduction of a damping load reduce the gain of the klystron, so that more driving power is required.

The output cavity is coupled to a 3 in-diameter coaxial feeder which can be connected by means of $\frac{1}{4}$-wave stub switches to a test load or via the combining unit to the aerial.

(4.1.3.1) *Methods of Broadbanding.*[12,13,14]

To obtain the required bandwidth and frequency response of the klystron amplifier the following experimental methods were tried:

(*a*) The use of a dissipative line between the modulated amplifier and the klystron input cavity employing sufficient coaxial r.f. cable type SAL14M (loss approximately 18 dB per 100 ft) to give about 6 dB insertion loss.
(*b*) Stagger-tuning of the centre cavity.
(*c*) Damping of the input cavity.
(*d*) Damping of the centre cavity.

Combinations of two or three of the above methods, e.g. (*a*), (*b*) and (*d*) were also used, as well as detuning of the output cavity from the carrier frequency towards the centre frequency of the band. This last method was abandoned as it resulted in a considerable loss of power. After carrying out detailed investigations and testing the frequency response at the klystron output, both by point-by-point measurements and by using a frequency sweep and a sideband analyser, it was found that best results were obtained by combining (*b*) and (*c*), i.e. by damping the input cavity to the extent of about 10 W by a coupled-in external load, and by detuning the centre cavity towards the sound carrier frequency. The frequency response shown in Fig. 5 for the 625-line system was obtained by these means. For 405-line operation, adequate frequency response was obtained by combining methods (*a*) and (*b*).

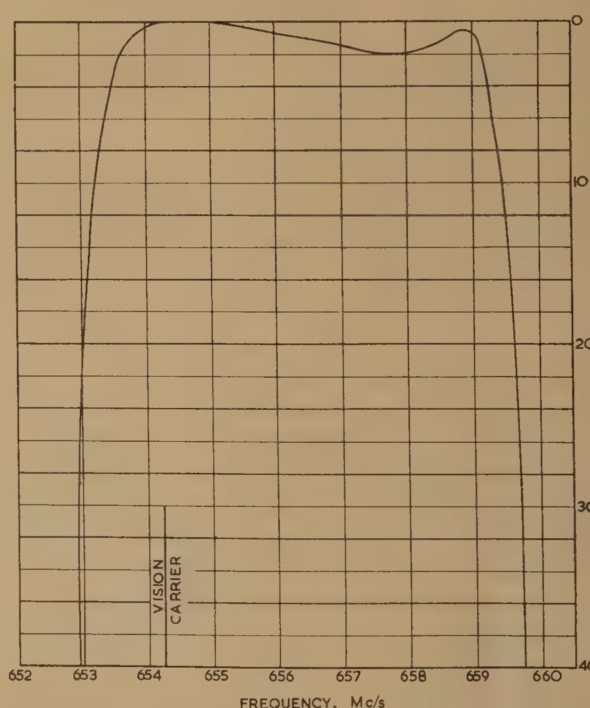

Fig. 5.—Overall frequency response (625-line system) at the output of the combining filter.

(4.1.4) **Klystron Cooling.**

Table 3 shows the cooling arrangements for the klystron.

Table 3

COOLING ARRANGEMENTS

	Cooling medium	Flow	Pressure drop
Collector	Water	20 g.p.m.	28 lb/in^2
Drift tubes (four in series)	Water	1 g.p.m.	6·4 lb/in^2
Electron gun envelope	Air	52 ft^3/min	5 in standard water gauge
Output cavity	Air	50 ft^3/min	1 in standard water gauge
Centre cavity	Air	20 ft^3/min	$\frac{1}{2}$ in standard water gauge

The water from the collector is cooled in a heat exchanger capable of dissipating 50 kW.

The air cooling of the centre cavity reduces tuning drift of this cavity during operation; it is not required merely for cooling its ceramic window.

(4.2) **Modulated Amplifier**

For narrow-band operation, a driving power of 25 watts is specified to produce an output of about 10 kW, but for a bandwidth appropriate to a vision transmitter it is necessary to stagger the tuning of the cavities and possibly also to damp them; this reduces the gain of the klystron, which then requires a driving power of the order of 100 watts to give the same power output.

In this experimental transmitter it was decided to provide a driving power of 200–300 watts. With this extra driving power available it becomes possible to tune the circuits to a wide stagger and to introduce damping loads to increase the bandwidth. To produce such driving power at frequencies up to 960 Mc/s, an experimental valve, type E2248, is used. This is a disc-seal tetrode, with a rated anode dissipation of 500 watts, in which great care has been taken to align the grid and screen wires accurately.

The E2248 is used as a modulated amplifier in the cathode-modulated cathode-driven circuit shown in Fig. 6. Published information on television transmitters constructed in this country shows that, so far, they have all been grid modulated.* The decision to use cathode modulation resulted from a consideration of the following advantages:

(*a*) It is possible to employ a construction with the grid decoupled to earth through a large mica capacitor, which leads to a greater stability throughout the whole band.

(*b*) A single h.t. supply can be used for both the modulated amplifier and the modulating valves. This supply can have its negative pole earthed and no floating supplies are required.

(*c*) The modulating valves feed into a more constant impedance.

A disadvantage of this arrangement is that the filament winding of the modulated amplifier is live at video frequency and a transformer having a low stray-capacitance filament winding is essential. A half-wave line is used to feed r.f. drive to the E2248 cathode, with the video signal connected at the r.f. nodal (quarter-wave) point.

The input and output circuits consist of adjustable coaxial cavities (Fig. 7). The output cavity operates in a quarter-wave mode and is tuned by a short-circuiting slider. For the highest frequency range of Band V, with the slider at the lowest position, further tuning is carried out by four diagonally arranged plungers which reduce the effective volume of the output cavity by sliding in horizontally towards the valve (this arrangement is not shown in Figs. 7 and 8). The valve is inserted through the 5 in-diameter inner tube, to which the anode is connected through a 1 000 pF capacitor. These capacitors provide r.f. by-pass between the valve and the cavity whilst insulating the cavity from high direct

* Cathode modulation has been used in sound broadcast transmitters.

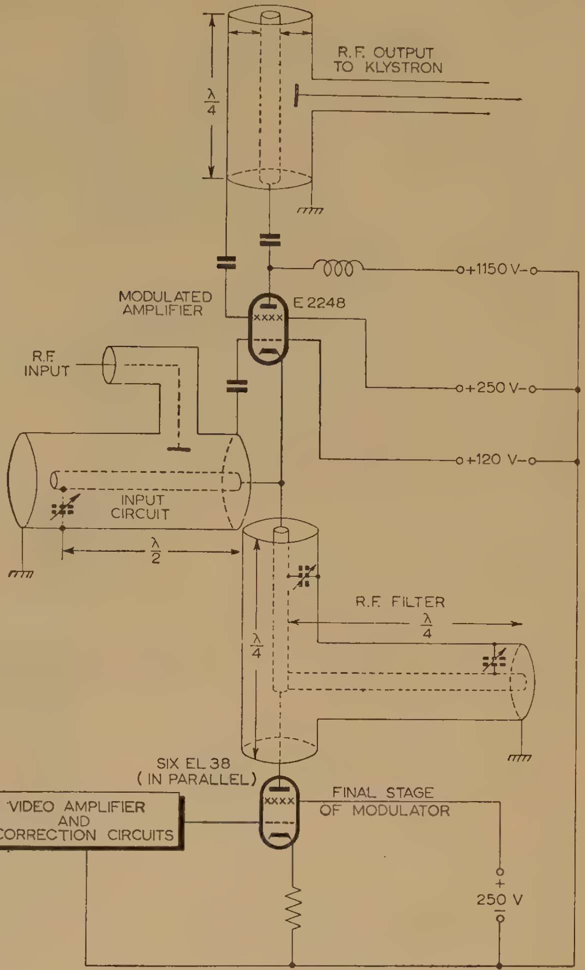

Fig. 6.—Modulated amplifier schematic.

voltage in the anode and screen-grid supplies. Both capacitors are mica-plate type, built into the valve holder (Fig. 8). Because of cathode modulation, the input cavity operates in a half-wave mode and is tuned by a telescopic length adjustment of the inner tube (the length of the outer is determined by the lowest frequency, i.e. 470 Mc/s, and is constant); in addition, fine tuning is provided by a small variable capacitor at the top of the cavity (i.e. near the valve at a point of fairly high r.f. potential).

The control grid is connected to the outer tube of the input circuit and the cathode to the inner, both through mica-plate capacitors built into the valve holder (similar to those in the output cavity). The r.f. driving power from the second 4X150G frequency doubler (see Section 4.4) is carried through a UR67 coaxial cable, and capacitively coupled to the inner of the input circuit; this coupling is adjustable. The filament lead and cathode connection are brought out through resonant lines at a point of low r.f. potential. The output power is carried either by capacitive coupling to, or by a direct tap on, the inner of the output cavity, depending on the frequency and mode of operation. The output power is carried by UR67 coaxial cable to the klystron input cavity. Neutralization was unnecessary with the precautions taken to ensure that there was no common path for the r.f. currents in the input and output cavities. The cavity between the grid and screen was arranged to resonate at as high a frequency as possible to prevent it resonating within the operating range.

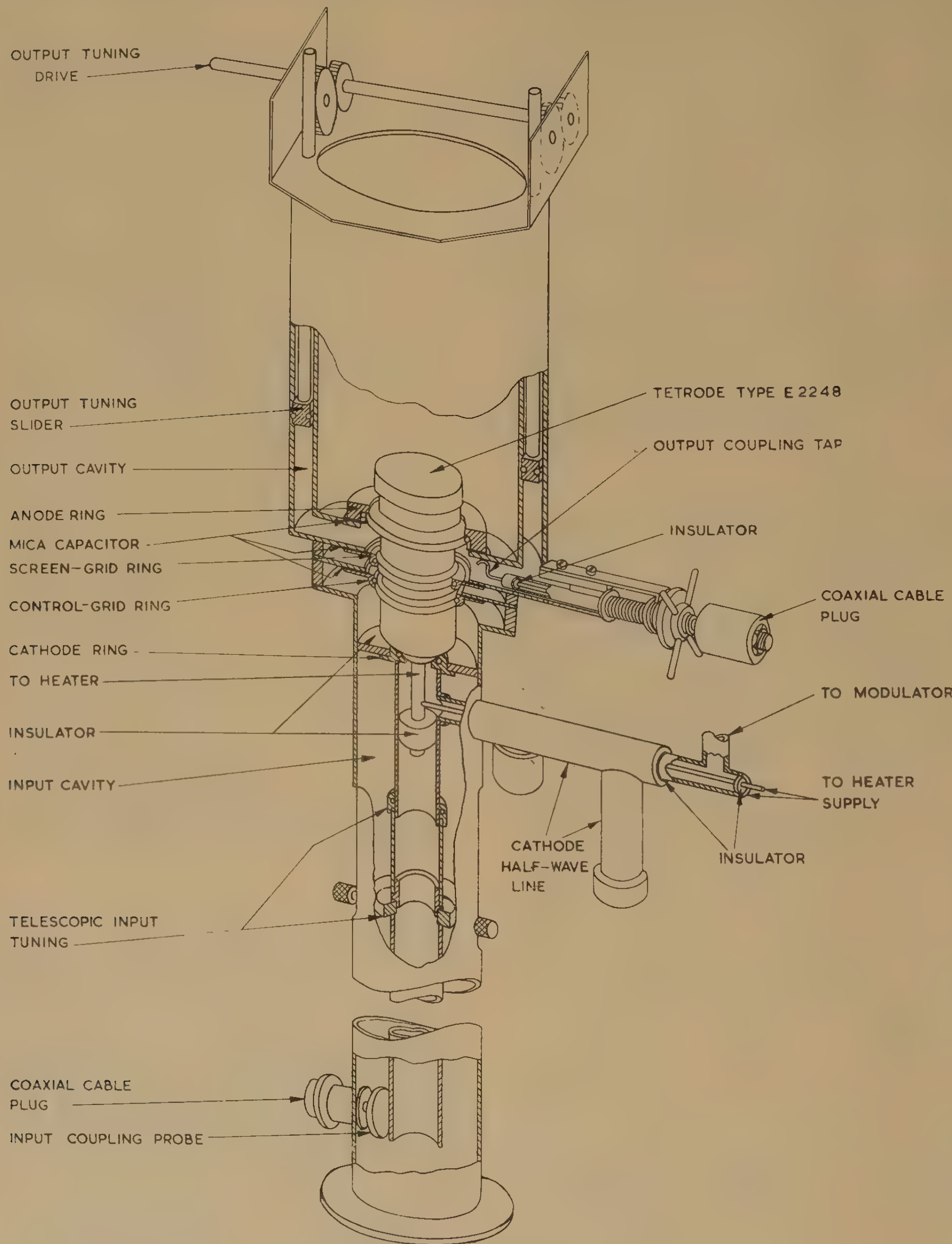

Fig. 7.—Modulated amplifier cavities.

(4.3) Crystal Oscillator Unit

The crystal oscillator unit is of a conventional design using a crystal of about 10 Mc/s (in a thermostatically controlled oven) in a feedback circuit, followed by frequency multipliers giving a total multiplication factor of 18.

(4.4) Intermediate Power R.F. Stages

Two frequency doublers follow the crystal oscillator unit, each using a 4X150G disc-seal u.h.f. tetrode capable of dissipating 150 watts. The input circuits, fully screened, consist of strip lines operating in the half-wave mode and driving the control grid, the remote ends of the lines being terminated by small fine-tuning capacitors. Each output circuit employs a coaxial cavity tuned to a quarter-wave mode by a slider. A 5 in-diameter inner was chosen for easy replacement of the valve, which i seated inside the cavity. The anode is connected to the inner and the screen grid to the outer, at r.f., by means of mica plate capacitors of about 1 000 pF, forming part of the valv holder. The cathode and screen grid are at r.f. earth potential Output coupling to the anode is through a capacitive probe (o alternatively a direct tap on the inner) variable in position alon the slot in the outer cylinder; the depth of penetration can als be varied to obtain optimum coupling. Power transfer t subsequent stages is through UR67 coaxial cable. As the inpu and output circuits are well screened from one another an operate at different frequencies, no neutralizing is required.

The valve is forced-air cooled and suitable for the whole c Bands IV and V. It is used self-biased by a grid resistor c about 25 kilohms and there is also a safety cathode-bias resisto

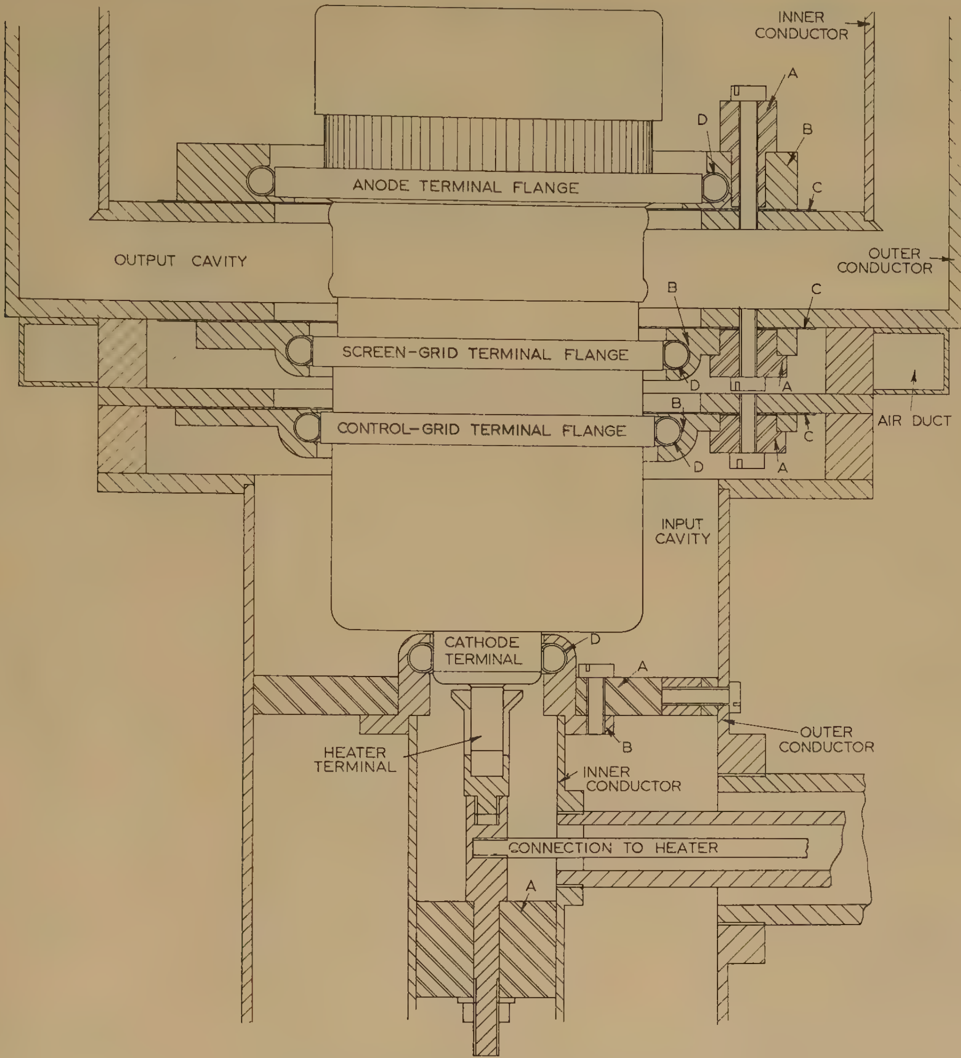

Fig. 8.—Tetrode type E2248 assembly.

A Insulator.
B Brass ring.
C Mica capacitor.
D Spiral contact spring.

which is brought into circuit by a relay operated by the grid current to protect the valve in the absence of drive. In the second frequency-doubler, the heater voltage is reduced by about 15% when operating under r.f. conditions to allow for transit-time back-heating effects at frequencies above 400 Mc/s.

(4.5) Modulator

The vision modulator uses receiving-type valves only. The whole circuit, consisting of amplifiers, clamp-pulse generators and waveform pre-distortion circuits, is contained in one unit. A schematic is shown in Fig. 9.

The output stage consists of six type EL38 valves connected in parallel. Individual cathode resistors improve linearity and balance the valve currents. This stage has adjustable bias applied through a clamp to set blanking level at the transmitter r.f. output independently of the levels in the remainder of the modulator.

Because of the high input capacitance, the output stage is driven by a cathode-follower, which, in turn, is fed from a paraphase amplifier consisting of a cathode-coupled pair of valves. The signal may be taken from either anode of this amplifier according to the sense of modulation required.

Between the linearity correction stage (marked 'Bend' in Fig. 9) and the paraphase amplifier there is a further stage of synchronizing pulse boost followed by a stage to limit the synchronizing pulses.

Clamping of suppression level is carried out in certain places in the modulator. A feedback type of clamp operates from the grid of the first valve to the output of the synchronizing pulse boost circuit.

Non-feedback-type clamps are provided at all other grids where clamping is necessary.

Under normal working conditions the clamps are in operation and d.c. bias can be applied to the various stages through the clamps. Under test bias conditions the clamps are automatically made inoperative by biasing potentials applied to the anodes and cathodes of the diode switches, and grid-bias potentials are applied through permanently connected grid-bias resistors. The

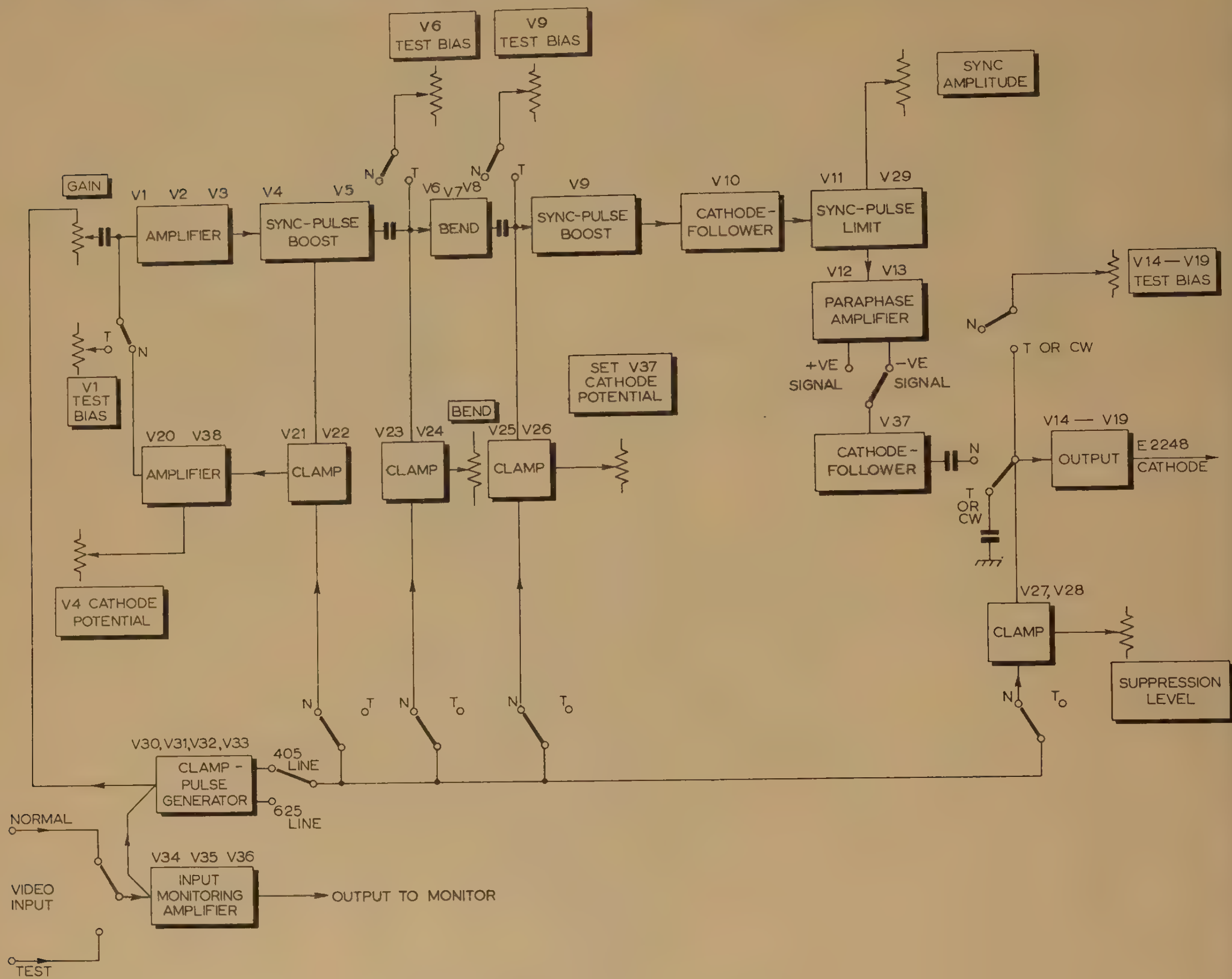

Fig. 9.—Vision modulator block diagram.

Switch markings: N Normal.
T Test bias.
CW Continuous wave.

linearity correction stage, the synchronizing boost amplifiers and the input amplifiers are all of conventional design.

(5) SOUND TRANSMITTER

(5.1) Low- and Intermediate-Power R.F. Stages

In the sound transmitter the low- and intermediate-power r.f. stages, as well as the modulated amplifier, are similar to the corresponding stages in the vision transmitter described above.

(5.2) Modulator

A three-stage modulator is used for amplitude modulation. A variable attenuator (0 to −5 dB) between the a.f. input and the input transformer adjusts the level fed to the grid of a Z727 pentode which is resistance-capacitance coupled to the grid of an EL38 pentode; this in turn drives eight EL38 valves in parallel forming the modulator proper and connected between the cathode of the E2248 modulated tetrode and earth.

To improve linearity, overall negative feedback of about 18 dB is applied by means of a rectifier probe in the output feeder from the klystron output to the first stage of the modulator.

(5.3) Power Amplifier

Both in mechanical design and in electrical circuit the klystro amplifier in the sound transmitter is almost identical with tl vision amplifier described above. The centre cavity is detun from resonance, as a precaution against over-voltage in tl cavity. The carrier power output is set to $2\frac{1}{2}$ kW, correspondi to a peak output of 10 kW at 100% modulation.

In the early development stage, some trouble was expe enced due to noise in the output signal (noticeable on t cathode-ray oscillograph) caused by the so-called 'multipact effect'[6, 15, 16, 17, 18] in the klystron; this effect is due to t resonance of slow secondary electrons moving across t klystron gaps between the cavity gap electrodes. The mul pactor effect was removed by careful adjustment of the focusi magnetic field.

(5.4) Frequency Modulation

For the 625-line tests to the C.C.I.R. standards, the sou transmitter was frequency modulated. For this purpose a f quency-modulated drive unit providing about 5 watts at appro mately 55 Mc/s was followed by a frequency-tripler stage feed

into the first 4X150G frequency-doubler, the output of the second 4X150G doubler driving the klystron amplifier directly. Under these conditions the full power output of 10 kW could be obtained, but as the vision-transmitter peak power was 10 kW, the sound transmitter output was set to 2 kW, in accordance with C.C.I.R. standard 5 : 1 vision/sound ratio. A much smaller sound transmitter can be used as companion to the 10 kW vision transmitter on the C.C.I.R. standard, and possible alternatives for the output amplifier are a 2 kW three-cavity klystron or a u.h.f. tetrode of similar power.

The above method of obtaining the u.h.f. frequency-modulated output with a deviation of ±50 kc/s is not ideal owing to the difficulty of achieving good noise figures with so many stages of frequency multiplication after the modulator. It was used in this instance as it enabled an existing v.h.f. drive unit to be used with the minimum of modifications to the transmitter.

tained in an oil-filled tank approximately 4 ft cube, with cooling-water coils in the top of the tank.

The bombarder supply provides 0·7 amp at 2·3 kV. The negative terminal of this supply is at 19·3 kV to earth, and it was therefore convenient to enclose both the transformer and the metal rectifiers in an oil-filled water-cooled tank forming a compact unit.

The filament of the klystron is operated at about 20 kV negative to earth, and the filament transformer is cast in resin to provide the necessary insulation for the secondary winding.

(7) PERFORMANCE

Table 4 gives the essential performance figures obtained at the final tests at Crystal Palace, and Table 5, some typical data of klystrons in varying operating modes.

Table 4

SUMMARY OF TESTS AT CRYSTAL PALACE

	British Standards	C.C.I.R. Standards
Vision transmitter	654·25 Mc/s	654·25 Mc/s
Modulation system	405-line positive amplitude modulation	625-line negative amplitude modulation
R.F. output power	10 kW c.w. at peak white level	10 kW c.w. at peak synchronizing pulse level
Total input power from mains	Transmitter approximately 45 kW Cooling equipment 5 kW	Transmitter approximately 45 kW Cooling equipment 5 kW
Frequency response at the output of the combining filter	Flat within ±0·5 dB to 2 Mc/s; ±1 dB to 2·5 Mc/s; −2 dB at 3 Mc/s; at −0·75 Mc/s −3 dB; 6 dB attenuation between the modulated amplifier and the input cavity; centre cavity stagger-tuned	Flat within ±0·5 dB to 2 Mc/s; ±1 dB to 4 Mc/s; −2 dB at 4·8 Mc/s; at −0·75 Mc/s −3 dB; Centre cavity stagger-tuned; input cavity damped
Transient response	Time of rise or fall 0·18 μs with 5% maximum overshoot. Ring frequency above 3 Mc/s	Time of rise or fall 0·16 μs. Ring frequency above 4·5 Mc/s
Sound transmitter	650·75 Mc/s	659·75 Mc/s
Modulation system	Amplitude modulation	Frequency modulation
R.F. output power	2·5 kW carrier, 10 kW peak	2 kW (set by lowering the beam voltage to 11·7 kV and driving the klystron directly by the second 4X150G frequency doubler)
Total input power from mains	Transmitter approximately 45 kW Cooling equipment 5 kW	Transmitter approximately 22 kW Cooling equipment 5 kW
Frequency deviation	—	50 kc/s (for 100% modulation)
Frequency response	±0·5 dB in the range 100–5 000 c/s and ±2 dB in the range 30–10 000 c/s at an input voltage equivalent to 70% modulation at 400 c/s	±0·5 dB between 60 and 10 000 c/s, falling to −1·3 dB at 15 000 c/s, at 50 kc/s deviation
Harmonic distortion	Not exceeding 1%, 1·5% and 4% at modulation levels of 30%, 70% and 90% modulation respectively	At 50 kc/s deviation, not exceeding 0·8% between 60 c/s and 5 000 c/s and up to 1·5% at 15 000 c/s
Noise level	50 dB below signal level at 90% modulation	−43 dB relative to 50 kc/s deviation at 1 000 c/s
Frequency stability	±5 parts in 10^6	±20 parts in 10^6

In an equipment specially designed for Bands IV and V, it would probably be better to obtain the output frequency by a heterodyne process to enable a better signal/noise ratio to be achieved.

(6) POWER SUPPLY EQUIPMENT

Selenium metal rectifiers have been used throughout the transmitters, as this type is more reliable than the valve type, although involving a rather greater initial cost. All the smaller metal rectifiers have been rated for ambient temperatures up to 25° C with occasional peaks of short duration up to 35° C without artificial cooling.

The high-power supply for the klystron beam h.t. has to deliver 2·0 amp at 17 kV. The rectifiers for this unit are con-

(8) TEST LOAD

It was decided to construct a test load in which the power-absorbing dielectric was the circulating water. In order to obtain a minimum reflection coefficient, it was necessary to design the loading line in such a way that the proportion of water dielectric increased along the length of the taper, away from the source, thus increasing the attenuation per unit length in the direction of power flow until all power was practically absorbed in the water. In view of the expense of machining tapers following a complicated law, it is fortunate that it was found possible to employ a linear taper giving a satisfactorily low reflection coefficient for frequencies above 470 Mc/s with a reasonable length of taper.

The water load arrangement finally adopted consisted of outer

Table 5

SOME TYPICAL KLYSTRON OPERATING DATA

	Vision transmitter	Sound transmitter*	
		a.m.	f.m.
Beam voltage, kV ..	16·6	17·1	11·7
Beam current, amp ..	1·8	1·66	0·8
Beam power input, kW ..	29·8	28·4	9·36
R.F. output power, kW ..	10	2·5 (carrier)	2
Beam efficiency, % ..	33·5	8·8 (carrier)	21·4
Body current, mA ..	60	66	32
Filament voltage, volts ..	7·25	8·8	8·2
Filament current, amp ..	33·5	43·3	40·5
Bombarder voltage, kV ..	2·25	2·15	1·84
Bombarder current, amp	0·64	0·72	0·74
Bombarder power, kW ..	1·44	1·55	1·36
Electrostatic focus, volts	−94	0	0
Electromagnetic focus:			
Coil No. 1, AT	540	450	345
Coil No. 2†	0	0	0
Coil No. 3, AT ..	4 050	3 700	3 750
Coil No. 4, AT ..	3 600	3 575	3 600
Coil No. 5, AT ..	2 150	1 640	1 820

* Older type klystron with higher filament voltage.
† Conforming to the initial instructions of the valve makers, all three main coils Nos. 2, 3 and 4 (also called body coils) were to be energized. Fig. 3 refers to this case. Later, a somewhat lower magnetic field was found adequate for the optimum performance and only body coils Nos. 3 and 4 were energized.

and inner conductors of constant diameter and a conically-shaped polythene insert which determined the proportion of water dielectric across the section of the line. The length of the taper was over two wavelengths in air for the lowest frequency of 470 Mc/s.

At the supply end the water load has a characteristic impedance $Z_0 = 51{\cdot}5$ ohms; along the load, Z_0 decreases continuously, owing to the increasing proportion of water dielectric. A voltage standing-wave ratio better than 1·1 : 1 was obtained on a line terminated by the load over the full frequency range. An s.w.r. of 1·02 : 1 was obtained over a narrow band of frequencies by means of fine-adjustment matching stubs in the feeder joining the klystron output to the test load.

(9) MECHANICAL DESIGN

The overall mechanical design of these transmitters has been given careful attention to ensure reliability, simplicity of operation, accessibility of all components and ease of manufacture.

The large transformers and rectifiers with their smoothing circuits are located in an interlocked enclosure behind the transmitter.

The layout adopted is shown in Fig. 10. The vision and sound transmitters are arranged in line. Access to the enclosure behind the cubicles is possible only through interlocked doors in the side screens. In this enclosed space are mounted the main beam h.t. transformer and rectifier tanks, the h.t. smoothing unit and the bombarder h.t. supply tank. By using water cooling for these oil-filled tanks it has been possible to reduce their size. Similar size units could have been used with air cooling, but it would have been necessary to provide ducts in the enclosed space for the air, and, in addition, a larger air blower would have been required.

The cubicles are made entirely of sheet steel folded for rigidity, and the whole cubicle is welded together to form a comparatively light unit. Sliding doors are used on all cubicles. These have the advantages that when the doors are open access to the adjacent cubicles is not impeded, and it is not necessary to allow

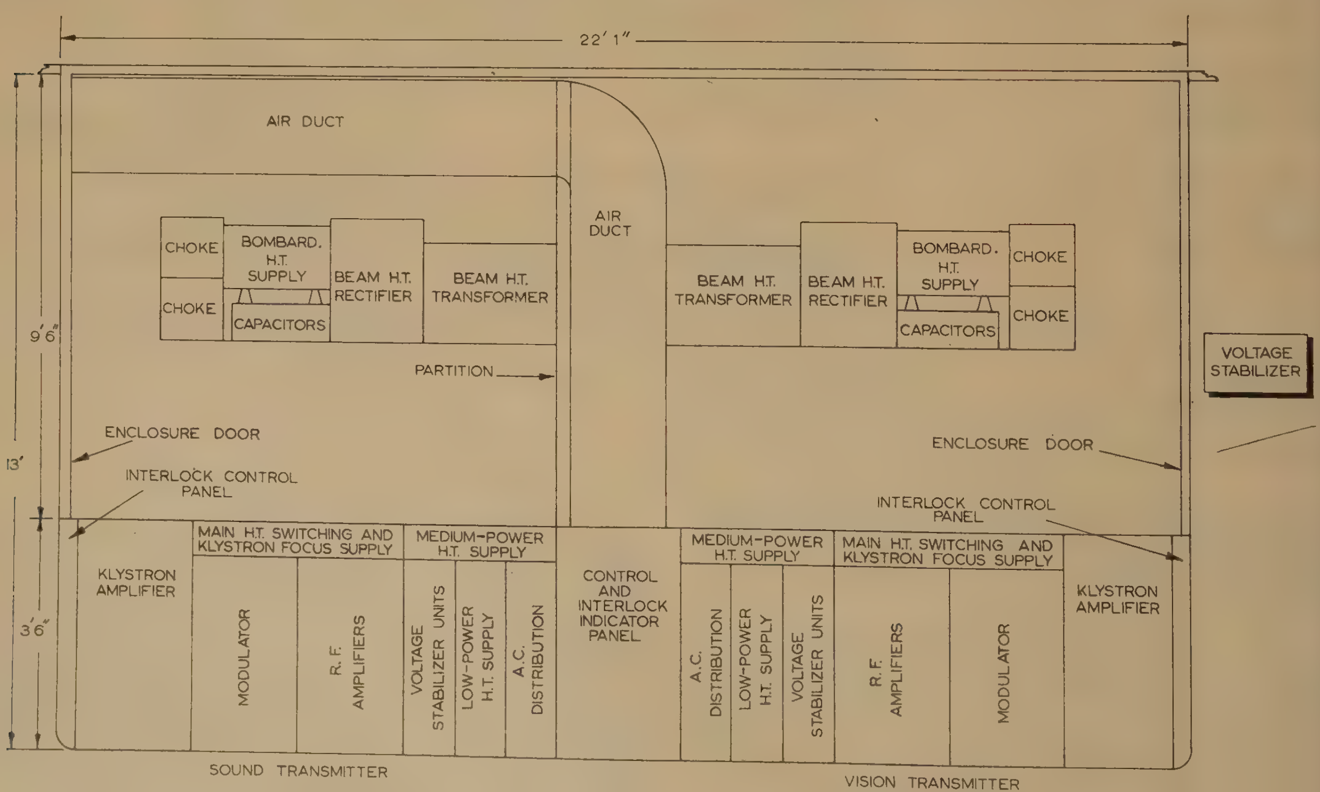

Fig. 10.—Layout of the transmitter (with vision and sound transmitters in line).

Height of cubicles 7¼ ft; head room of 2 ft required above cubicles.

space for the doors to swing and therefore the overall floor area required for the transmitter could, if necessary, be reduced. Where space is very limited, the vision and sound transmitters can be mounted at right angles to each other.

The majority of the units mounted in the cubicles are fitted in sliding racks that can be fully withdrawn to give access to components. An air duct runs along the base of all cubicles, through which air is blown into each rack of equipment. All inter-cubicle wiring is carried in the roof space behind the meters.

A comprehensive mechanical interlock system is employed which operates in three phases to ensure that the interlock circuits are broken, the main supply is isolated and the h.t. supply is earthed before the doors or sliding panels can be opened. The mechanical interlock is unique in one important aspect. It has been accepted that it is an advantage, during periods of testing, to operate the transmitters with the interlocks out of action and, under proper supervision, to apply power to the transmitters with the doors and racks open. A key is therefore provided which can be used to isolate the door-locking system. The turning of this key also switches on red flashing lights in the cubicles and in the enclosures to indicate that the transmitters are being operated in an unsafe condition. Provided that none but responsible engineers are given access to this key, and that it is used only with proper supervision, its use constitutes no greater danger to staff, whereas it considerably accelerates the adjustment of the apparatus.

(10) CONCLUSIONS

Using a commercially available klystron valve, a successful vision transmitter with a peak output power of 10 kW operating at about 650 Mc/s, together with its accompanying sound transmitter, has been made. These transmitters fulfil all the usual requirements of the specifications for 405-line or 625-line vision transmitters and for high-quality sound broadcast transmitters (a.m. or f.m.).

The transmitters are straightforward to set up and are easy to operate. Vision transmitter tests can be greatly facilitated by the use of a sideband analyser giving a swept-frequency display of both upper and lower sidebands. In comparison with transmitters of equivalent power at lower frequencies the efficiency is lower, but it is no less than can be tolerated for the economic operation of television transmitters of such power.

For the successful exploitation of the u.h.f. channels for television broadcasting it will be necessary to construct transmitters of considerably greater power than those described in the paper. It is expected that this greater power can be produced by larger klystrons, probably of the four- or five-cavity type. This type would make further stagger tuning possible and would have an inherently greater gain, so that adequate bandwidth should be obtainable with driving powers of the order described, possibly even less. In the case of a higher-power and higher-gain klystron with external cavities, extra care will be necessary to ensure good contact between the cavities and the klystron, otherwise instability may occur due to feedback caused by leakage from the output cavity. The practical difficulty in accomplishing this may prove an argument in favour of the integral-cavity type of klystron.

There is no reason to doubt the suitability of wide-band multi-cavity amplifiers for future use in colour television. In particular, the differential phase/frequency response should be adequate.[13]

Although the description of the transmitters given in the paper relates to their use for television broadcasting, it is obvious that their application is not limited solely to this field, and the vision transmitter could readily be adapted for any service requiring wide-band operation at ultra-high frequencies, e.g. for multi-channel communication utilizing tropospheric scatter propagation.

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THE FIELD STRENGTHS REQUIRED FOR THE RECEPTION OF TELEVISION IN BANDS I, III, IV AND V

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(*The paper was first received* 20*th June, and in revised form* 5*th November*, 1958. *It was published in January*, 1959, *and was read before the* ELECTRONICS AND COMMUNICATIONS SECTION 27*th April*, 1959.)

SUMMARY

The paper examines the factors involved in the reception of 405-line television signals in the four frequency bands assigned by international agreement to television broadcasting in Europe. After consideration of the receiving-installation characteristics appropriate to the various bands, and the statistical variation of field strength with aerial location over the range of frequencies involved, a basis for the specification of the median field strength defining the nominal limit of the service area of a television station is described.

The estimates made, particularly for Bands IV and V, must of necessity be provisional: it may be necessary to modify them when first-hand experience of operation in these Bands becomes available. Despite this limitation, however, it is considered that the figures given provide a useful interim guide for planning purposes.

(1) INTRODUCTION

Television broadcasting in Europe is allotted four separate frequency bands, in accordance with the international frequency-allocation plan formulated at the Atlantic City Radio Conference in 1947, the designations and limits of these bands being

Band I	41–68 Mc/s
Band III	174–216 Mc/s
Band IV	470–585 Mc/s
Band V	610–960 Mc/s

Band II (88–100 Mc/s) is assigned to sound broadcasting and is therefore not considered here.

In this study, consideration is given to the lower limit of field strength at which 405-line reception with modern receiving installations provides a signal/noise ratio at the picture tube meeting a definite subjective standard, when this limit is set solely by natural and receiver noise. These circumstances are usually described as 'rural' reception, for in 'urban' reception the minimum field strength required is set largely by man-made noise, mainly impulsive in character. The effect of man-made interference is a complex matter requiring separate examination, for which adequate statistical information is not yet available, especially in the higher bands. A further question not considered in the paper is the degradation of picture quality caused by multi-path propagation.

In defining the standard of signal/noise ratio which forms the basis of the paper, use has been made of information submitted to the appropriate study groups of the International Radio Consultative Committee (C.C.I.R.) by the United Kingdom. This specifies the signal/noise ratio at the modulating electrode of a picture tube giving a television picture of a desirable subjective standard. By working back through the stages of a typical asymmetric-sideband receiver, the corresponding signal/noise ratio in the early stages of the receiver can be derived. This derivation is carried out in Section 2, and results in an r.m.s. peak-white signal/r.m.s.-noise ratio of 36 dB for what is termed 'satisfactory' reception.

It is then necessary to determine for each frequency band the field strengths which, with practical receiving installations, will provide this signal/noise ratio in a rural location. In order to arrive at median field strengths which can be regarded as the nominal limits of 'satisfactory' reception, consideration must be given to the variation—which in certain circumstances may be quite large—of field strength with aerial location within a small area. This leads to an examination of the local variation factor appropriate to each of the bands and finally to a suggested basis for defining the nominal limits of the service area of a television station.

Mr. Swann is in the Post Office Engineering Department.

(2) FIELD STRENGTH REQUIRED AT AN AERIAL

(2.1) Signal/Noise Ratio Required at the Receiver Input for 'Satisfactory' Reception

The minimum acceptable signal/noise ratio at the receiver input is determined, basically, by that value of video signal/noise ratio at which the noise just begins to degrade the observed picture. Subjective tests over a long period carried out for the C.C.I.R.* have resulted in the adoption of the figure of 12 dB as the acceptable minimum ratio between peak-to-peak amplitude of the picture signal (excluding synchronizing pulses) and that of random noise. Expressed in the form more recently proposed to the C.C.I.R., the ratio between the peak-to-peak picture signal and the r.m.s. noise should be 30 dB; this allows a crest factor of 18 dB to convert the r.m.s. amplitude of random noise to its peak-to-peak value.†

For simplicity, we first consider conventional double-sideband reception using a linear, or envelope-following, detector. In this case, and ignoring a constant factor of proportionality involved in linear detection, the peak-to-peak video signal out of the detector is equal to the peak r.f. signal input (at peak white). Since the picture content of the output signal is only 70% of the total video amplitude, the peak-to-peak picture-signal output is equal to the r.m.s. r.f. signal input (at peak white). Furthermore, the r.m.s. amplitude of noise is unaffected by the process of linear detection if we again ignore the proportionality factor common to both signal and noise and if the noise amplitude is small compared with the signal.† In other words, the r.m.s. noise output at video frequency is equal to the r.m.s. noise input measured at radio frequency. We may thus conclude that the ratio between the peak-to-peak picture signal and the r.m.s. video noise has the required value of 30 dB if the input signal/noise ratio, expressed as the ratio between the r.m.s. value of r.f. signal at peak white and that of the r.f. noise, is also 30 dB. The values of r.f. signal and noise are, of course, here measured after those early stages of the receiver in which some additional noise is introduced, and the noise bandwidth is limited to that which will produce effective noise in the receiver output.

It now remains necessary only to allow for the relative effects upon signal and noise of asymmetric-sideband reception. So far

* C.C.I.R.: 'Documents of the VIII Plenary Assembly, Warsaw, 1956', Volume 1, Recommendation No. 154, p. 121.

† GOLDMAN, S.: 'Frequency Analysis, Modulation and Noise' (McGraw-Hill, 1948), p. 244.

as the signal is concerned, the effect is to reduce its level by 6 dB. This is because of the cut in i.f. response, to half-amplitude, at the vision carrier and at nearby sideband frequencies and also of the virtual suppression of one sideband of all high-frequency components of modulation. The reduction in total effective i.f. bandwidth also reduces the noise, but no factor is introduced here for this, since, in the following Section, only the resultant effective noise bandwidth of 2·75 Mc/s is assumed in deriving the basic level of noise. With this provision, therefore, we may state that the required ratio of r.m.s. carrier (at peak white) to r.m.s. noise in the early stages of the receiver is 30 + 6 = 36 dB.

This result, it is emphasized, has been derived from a C.C.I.R. figure of 12 dB expressing the acceptable ratio of peak-to-peak picture-signal amplitude to peak-to-peak noise amplitude at the cathode-ray tube. While it is possible that some slight modification may be found desirable because of the increased size and brightness of modern tubes, insufficient information is available at present to justify any departure from this value of 12 dB. Throughout the paper, reception at this signal/noise ratio is arbitrarily called 'satisfactory', although it is emphasized that —depending upon circumstances such as the type and interest value of the actual picture—an appreciably lower ratio can be quite acceptable to many viewers. In the planning of television services, however, a single standard of reception must be adopted which does not depend upon such circumstances for its acceptability.

(2.2) Field Strength Required at a Dipole to give 36 dB Signal/Noise Ratio in an Ideal Receiving Installation

In a noiseless receiver which does not load the aerial and which has no feeder losses, an output signal/noise ratio of 36 dB will be developed when the ratio between the signal and the noise e.m.f.'s in the aerial is 36 dB. A typical 405-line receiver has an energy bandwidth of 2·75 Mc/s, and the thermal-noise e.m.f. available in the nominal 75 ohms internal impedance of a dipole for this bandwidth can be found from

$$V^2 = 4kTRf$$

where K = Boltzmann's constant.
T = Absolute temperature, deg K.
R = Aerial impedance, ohms.
f = Energy bandwidth, c/s.

The noise e.m.f. is found to be $1\cdot82\,\mu$V, or approximately 5·5 dB relative to $1\,\mu$V. It therefore follows that the signal e.m.f. must be 5·5 + 36, or 41·5 dB, relative to $1\,\mu$V. The field strength required to give this e.m.f. in a half-wave dipole at any frequency can be found by making use of the multiplying factor $\pi f/300$ (or π/λ), where λ is the wavelength in metres and f is the frequency in megacycles per second. This leads to the field strengths given in Table 1 for frequencies near the centres of the various bands.

The field strengths given in Table 1 are, of course, for an ideal

Table 1

FIELD STRENGTHS REQUIRED TO GIVE A SIGNAL/NOISE RATIO OF 36 dB IN A DIPOLE

Band	Frequency	Field strength required at a dipole aerial
	Mc/s	dB relative to $1\,V\mu$/m
I	55	36·5
III	200	47·5
IV	500	56
V	800	60

receiver of unity noise factor, with no feeder losses. They must be increased by the effective receiver noise factor and the feeder loss, and diminished by the gain, relative to a half-wave dipole, of the receiving aerial, to give the field strength required to provide a signal/noise ratio of 36 dB in a practical receiving installation.

(2.3) Characteristics of Typical Receiving Installations

(2.3.1) Receiver Noise Factors.

Some distinction should be made between the noise factor of a receiver alone and its effective noise factor when connected to an aerial. The general level of random background noise observed in practice is appreciably greater than thermal noise, and this is probably largely due to cosmic noise. This results in an effective noise factor higher than that measured in the laboratory, at least in the lower channels of Band I.

Published figures of noise factor for current designs of Band I and Band III receivers, and for sets designed for Bands IV and V, suggest that the values stated in Table 2 will be typical of sets used in areas close to the nominal limits of service areas.

Table 2

RECEIVER NOISE FACTOR

Band	Frequency	Noise factor		Effective noise factor (allowing for cosmic and man-made noise)	
	Mc/s	dB		dB	
I	55	6		8	
III	200	8		8	
		Type 1	Type 2	Type 1	Type 2
IV	500	14	8	14	8
V	800	16	10	16	10

Note: Type 1 refers to receivers using crystal mixers without pre-amplification.
Type 2 refers to receivers using low-noise r.f. amplifying stages which it is envisaged will be developed in the future.

Some comment on the noise factors of Band IV and V receivers is perhaps necessary. While receiver noise factors of 14–16 dB have already been achieved using turret-tuner inserts incorporating crystal mixers only and without preceding r.f. amplifiers, it is expected that noise factors of 8–10 dB will be attained using special r.f. valves. Currently available valves, even if of the ceramic-planar-triode type, give a noise factor of some 2 dB worse than the figures quoted for Type 2 receivers. The figures for both types of receiver are carried into the later conclusions of the paper, but in reading these it must be borne in mind that the prediction of future receiver noise factors is uncertain and the actual performance of domestic receivers will depend on economic factors such as the cost of providing a low-noise r.f. stage.

(2.3.2) Aerial Gains.

The decrease in size of a half-wave dipole at the higher frequencies and therefore the reduced signal level it picks up in the higher bands is only partially offset by the increased gain obtainable from aerials of practical and economic design. In addition aerials may fail to develop their plane-wave gain in many locations in built-up areas, although adequate statistical evidence of this effect is not yet available. Much more evidence is required before fully appropriate median values of aerial gain can be given for the different bands. Furthermore, a small loss should be allowed for possible mismatching. A conservative estimate should therefore be made of the aerial performance likely to b

Table 3

AERIAL GAINS AND FEEDER LOSS

Frequency	*Median gain, relative to half-wave dipole*	*Feeder loss*
Mc/s	dB	dB
55	3	1
200	6	2
500	9	3
800	11	4

obtained, particularly in Bands IV and V, and Table 3 gives, on this basis, median aerial gains which are thought to be appropriate for planning purposes. No allowance has been made for the possibility that full gains may not in all cases be achieved, owing to the need to deploy the directional characteristics of a receiving aerial for the reduction of echoes.

(2.3.3) Feeder Losses.

The feeder losses for the average domestic installation, using some 40–50 ft of cable, are about 1 dB for Band I, rising to perhaps 3 dB in Band III if solid-dielectric cable is used. Currently available air-spaced and cellular-dielectric cable can reduce the Band III loss to 2 dB, and at 800 Mc/s (Band V) the loss with such cable need not exceed 3 dB. It therefore appears that feeder losses will not have the importance in Bands IV and V originally expected, and values which now seem appropriate for planning are given in Table 3.

(2.4) Field Strength Required at a Rural Location with Practical Receiving Installations

Using the field strengths derived in Section 2.2 and the noise factors, aerial gains and feeder losses suggested in Tables 2–4, the field strengths necessary at a receiving installation in rural locations are derived in Table 4.

Table 4

FIELD STRENGTH REQUIRED AT RURAL LOCATION

Frequency	Field strength at a dipole to give a signal/noise ratio of 36 dB, with no feeder loss and a receiver of unity noise factor (see Section 2)	Allowance for noise factor of receiver (see Table 2)	Allowance for aerial gain (see Table 3)	Allowance for feeder loss (see Table 4)	Field strength required for 'rural' reception
Mc/s	dB rel. to 1 μV/m	dB	dB	dB	dB rel. to 1 μV/m
55	36·5	+8	−3	+1	42·5
200	47·5	+8	−6	+2	51·5
500	56·0	+14*	−9	+3	64·0*
500	56·0	+8†	−9	+3	58·0†
800	60·0	+16*	−11	+4	69·0*
800	60·0	+10†	−11	+4	63·0†

* Type 1 receiver. † Type 2 receiver.

For reasons discussed below, the field strengths given in Table 4 cannot be immediately accepted as the median values required for rural reception.

(3) FIELD-STRENGTH CONTOURS DEFINING THE NOMINAL LIMITS OF SERVICE AREA FOR EACH FREQUENCY BAND

Consideration of the field strengths required has been in terms of the actual field existing at the aerial, and no allowance has so far been made for the variation which occurs from point to point within a small area. This variation, which may be quite large, can be conveniently expressed by the 'local variation factor', which may be defined as the ratio, in decibels, between the field strengths exceeded at 50% and at 90% of the possible aerial locations within an area about one kilometre square. This factor will generally increase with frequency, but it is also influenced a great deal by the local terrain. For example, a factor of only 2 dB might be measured in flat rural areas in Band III, but this might rise to 12 dB or more in rough, heavily built-up areas. Fig. 1 shows the variation of this factor with

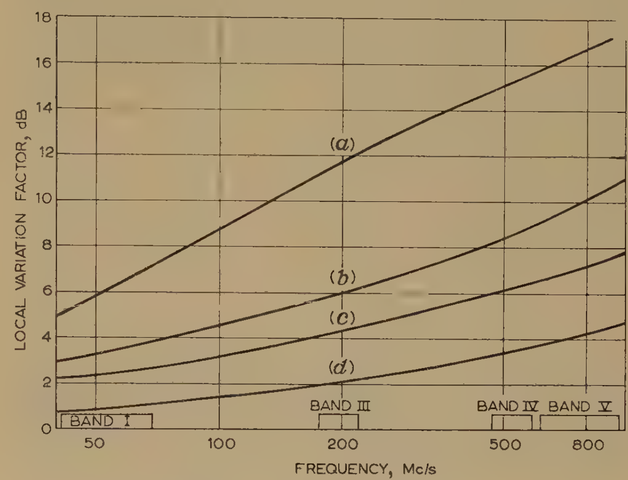

Fig. 1.—Local variation factor.

Ratio of field strengths exceeded at 50 and 90% of possible aerial locations.
(*a*) Moderately undulating built-up areas.
(*b*) Flat built-up areas.
(*c*) Moderately undulating rural areas.
(*d*) Flat rural areas.

frequency and terrain, but values for Bands IV and V must be regarded as tentative, in view of the few data so far accumulated.

Up to the present, field-strength contours defining the nominal limits of service areas have been in terms of median field strengths; this means that, near the contour, only 50% of the viewers receive a 'satisfactory' service. It would clearly be possible to draw contours such, for example, that 70 or 90% of the viewers within them received a specified signal/noise ratio or better. If the 90% criterion were to be adopted, this would merely mean using median field-strength contours for the field strengths given in Table 4 increased by the local variation factor appropriate to the frequency band concerned. However, this would mean that very many viewers outside the contour would be able to receive a 'satisfactory' service; to include in the service area only those viewers having a 90% probability of 'satisfactory' reception would therefore seem unrealistic. It must also be remembered that measurements of local variation have been made with a completely random series of aerial positions, whereas viewers' aerials are in practice often moved to obtain optimum results. Such adjustments are likely to be more effective in the higher bands than in Band I, and it can safely be assumed that more attention will be paid to aerial adjustment in these bands.

In consequence, the full effects of local variation will not be felt, and it is suggested that the adoption of contours giving a 70% probability of 'satisfactory' service on a random-aerial-position basis provides a useful compromise. The required median field-strength contours can be found by making an allowance for the ratio between the field strength exceeded at 50% of aerial locations and that exceeded at 70%, and this allowance can readily be derived from the values given in Fig. 1 if the distributions are assumed to be log-normal. That this is reasonable is illustrated by Fig. 2, which shows a typical distribution measured

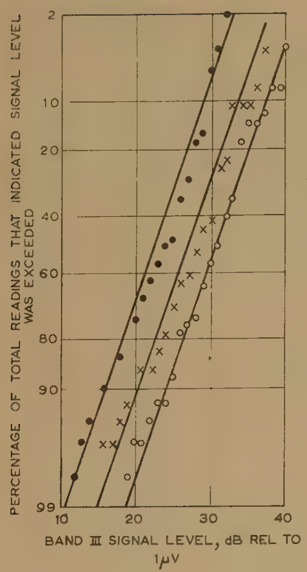

Fig. 2.—Signal-level distribution among 2-storey large semi-detached houses.

● ● ● Aerial height 12 ft 6 in.
× × × Aerial height 17 ft 6 in.
○ ○ ○ Aerial height 22 ft 6 in.

in a residential area in Band III. The appropriate allowance is made in Table 5, and the median field strength contours which will provide a 'satisfactory' service to 70% of the viewers at any rural location along them are thus obtained.

Table 5

MEDIAN FIELD STRENGTHS NECESSARY TO ENSURE AT LEAST 70% PROBABILITY OF SERVICE (RURAL AREAS)

Frequency	Field strength required for reception (in the presence of thermal and receiver noise only)	Local variation factor correction (50–70%)	Median field-strength contour to give 70% probability of service	
Mc/s	dB rel. to 1 µV/m	dB	dB rel. to 1 µV/m	mV/m
55	42	1	43	0·14
200	51	2	53	0·42
500	64*	2·5	66·5	2·0
500	58†	2·5	60·5	1·0
800	69*	3	72	4·0
800	63†	3	66	2·0

* Type 1 receiver. † Type 2 receiver.

The local variation factor for moderately undulating rural areas has been used in Table 5, and both Type 1 (high-noise) and Type 2 (low-noise) receivers are considered.

(4) SUMMARY OF FIELD-STRENGTH REQUIREMENTS

It is thus possible to summarize the minimum median field strengths required to provide at least 70% probability of service to viewers in rural areas. It is emphasized that this is on the basis of 'satisfactory' pictures being obtained when the ratio between the r.m.s. peak-white signal and the r.m.s. noise is 36 dB, although many viewers, at the present stage of television development, are reasonably satisfied with a somewhat lower ratio. While, therefore, a 36 dB signal/noise ratio leads to the nominal limits of service given below, pictures of a quality at present acceptable to many viewers will be obtainable outside these limits. It is, in fact, current practice for the broadcasting authorities in the United Kingdom to publish field-strength maps which show median field strengths down to 100 and 250 µV/m in respect of all channels in Bands I and III respectively.

Band I.—A field strength of 100 µV/m for Channel 1 and 150 µV/m for Channel 5 are adequate. Local variation effects are small in this band, and only 1 dB need be added to these values to define the median field-strength contours giving the nominal limit of service area.

Band III.—Reception necessitates a field strength of about 350 µV/m. A small local variation effect leads to a suggested median-field-strength contour for the nominal limit of service area of 450 µV/m.

Band IV.—Reception is limited by receiver noise at a field strength of about 1·6 mV/m when the simpler high-noise receivers are used, and a small local variation effect leads to a suggested service-area limit of 2 mV/m median field strength. The use of more expensive low-noise receivers would give an equal grade of service to the 1 mV/m contour.

Band V.—A field strength of about 2·8 mV/m is required when simple high-noise receivers are used. The local variation effect suggests that the service area should be defined by the 4 mV/m median contour. The use of low-noise receivers would make possible the extension of the nominal limit of service area to the 2 mV/m contour.

Table 6

SUMMARY OF FIELD-STRENGTH REQUIREMENTS

Band	Field strength required for 'just satisfactory' service			
	At individual receiving aerial location		To define contour limit of service area (70% probability of service)	
	mV/m		mV/m	
I	0·1–0·15		0·115–0·170	
III	0·35		0·45	
	Type 1	Type 2	Type 1	Type 2
IV	1·6	0·8	2	1
V	2·8	1·4	4	2

(5) CONCLUSIONS

An attempt has been made to determine the median-field-strength contours defining the nominal limit of service area for television reception in rural areas in Bands I, III, IV and V. The values arrived at for Bands I and III are very similar to those which have already been adopted in this country; those for Bands IV and V are necessarily tentative at this stage.

Rural reception in Bands IV and V should, however, be possible within the limits stated, since the basic data on receiver noise factor, aerial gain and local variation factor are unlikely to be seriously in error.

It should perhaps be added that the field strengths finally derived in Table 6 are not necessarily those at which protection from co-channel interference should be provided. For deducing these latter values, fresh consideration must be given to the percentage of viewers at a service-area fringe for whom protection should be planned, and this, in turn, demands some consideration of the space and time correlations between wanted and unwanted signal amplitudes.

(6) ACKNOWLEDGMENTS

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DISCUSSION BEFORE THE ELECTRONICS AND COMMUNICATIONS SECTION, 27TH APRIL, 1959

Mr. R. N. Jackson: Section 2.1 deals with a crucial point, namely the estimation of the signal/noise ratio required for satisfactory pictures. We have carried out a number of subjective tests in the laboratory in which varying amounts of random noise were presented in television pictures, and a team of observers were asked to assess the effects of this noise under carefully controlled conditions. A 21 in 405-line receiver was used, and various viewing distances from 3 ft to 7 ft 6 in and ambient light from almost zero to 5 ft-lamberts were investigated. Assessments were made in terms of six predetermined comments as used by the B.B.C. and others in recent field trials. These were (*a*) not perceptible, (*b*) just perceptible, (*c*) perceptible but not objectionable, (*d*) somewhat objectionable, (*e*) definitely objectionable, and (*f*) unusable. The results showed that noise in a 21 in receiver viewed at 3 ft under conditions of near darkness is just perceptible when the ratio of the peak-to-peak picture signal to peak-to-peak noise at the cathode is 22·5 dB and just objectionable when this ratio is 10·5 dB.

The significant fact uncovered is that these values vary with viewing conditions. Thus, when the ambient light is increased from almost zero (0·15 ft-lambert) to 3 ft-lamberts, the just-perceptible ratio is reduced to 17·5 dB. Further reduction occurs as viewing distance is increased.

The 'acceptable' signal/noise ratio depends upon our choosing some criterion for acceptability, but in the light of this evidence I would say that the figure of 12 dB upon which the paper is based is a reasonable one for a 21 in receiver, assuming that the viewer will usually have some lights on and will view from about 6–8 ft or more.

Mr. W. N. Anderson: The paper is based on the acceptability of a 12 dB limit, and at this separation of signal and noise, 50% of the observers will find reception satisfactory. The 1956 C.C.I.R. document is roughly in agreement on this point with document 19–E submitted to the C.M.T.T., 1958 Monte Carlo meeting. Has this figure of 50% satisfaction any effect on the 70% contour at the end of the paper?

Secondly, I feel that it may be necessary to consider the contribution of noise arising from the studio source in conjunction with the distribution network of lines. In order to study this we recently set up a 21 in picture display with a 12 dB separation between picture and noise and we played over it some typical video tape recordings. Three different sources from various parts of the country were tried, and on one we noticed that it was possible to see a definite picture impairment on account of noise. On the other two recordings there was no difference. It seems possible that, on the longest combination of links in the country and with the very widespread use of video recordings, there may be a significant noise contribution which must be added to the receiver noise considered in the paper.

The 50% probability contour, together with the 10% and 90% values, gives something which is just as useful and perhaps more readily understood than the 70% contour.

In terms of the number of people influenced by a broadcast, it is known, for example, that advertisements radiated in areas where the receiver density is as low as 10% have a very significant effect on sales in that area.

Mr. G. B. Townsend: I should like to draw attention to what I feel is a common experience of those who have travelled around the u.h.f. field in a mobile laboratory, namely the magnitude of the site-variability factor. At one site, which was only 15 miles from the transmitter, we could get an acceptable picture only with the aerial lying on the ground; in another, with a directional aerial facing more than half a mile of open water, there were 20 dB fluctuations in signal strength. This may well give rise to a serious problem in public relations; it is not always easy, and necessarily costs money, to find alternative aerial positions on a roof-top.

Regarding the 12 dB standard of satisfactory signal/noise ratio, I think it helps to overcome the differences due to experimental arrangements if for any particular experiment the minimum perceptible level and the just intolerable level are also quoted. Furthermore, does this figure refer to a receiver with a clean noise-free synchronization signal?

The author mentions the absence of information on impulsive noise in television pictures. Our experience is that if the frequency of occurrence of the impulses is fast enough for the observer to see several pulses at a glance, there is a difference of about 9 dB between the levels of impulsive noise and random noise which he can just see. The author suggests that a noise factor of 8 dB or better might be expected using r.f. amplifiers; I should be interested to learn other people's reactions to this, because I feel that we may only achieve reliably an average of 11 dB in our commercial receivers.

When discussing the use of the u.h.f. band it is tempting to turn to things which have yet to be decided, but if we did have a 6 Mc/s video bandwidth, how would this affect the author's figures? Would he merely add 3 dB? If we go to negative instead of positive modulation, would he make any change? Perhaps we ought to be thinking in terms of colour television. It is well known that in a true constant-luminance system the chromaticity changes caused by noise are some 8 dB less visible than the corresponding luminance fluctuations. If the noise level is appreciable, however, the chromaticity fluctuations are noticeable, and for an N.T.S.C. system I suggest that a further allowance of 2 dB should be added to the author's figures for this case.

Mr. A. B. Howe: The author derives theoretical figures for the field strength necessary to give satisfactory pictures in the various television bands. Recently, the B.B.C., with the co-operation of the G.P.O., D.S.I.R., I.T.A. and B.R.E.M.A., attempted to gain similar information by direct experiment. A 405-line television transmission was undertaken from the Crystal Palace site on 654 Mc/s with 125 kW e.r.p., and five mobile laboratories made a survey of the quality of the television pictures. Observations were also made of the pictures received from the B.B.C. television transmitter operating in Band I on 45 Mc/s. A subjective scale of six grades was used for the assessment of the pictures, and in assessing the results it was considered that if, at a given field strength, 70% of the observations were in the three more favourable grades, a satisfactory picture had been achieved.

The results obtained may be compared with those given in Table 4. Using the same method of assessment, but inserting the values for noise factor, etc., appropriate to the conditions of experiment, we arrive at a figure of 63 dB relative to 1 μV/m for the field strength necessary for satisfactory pictures. It is no doubt partly coincidence that the practical experiments also gave a figure of 63 dB, i.e. 1·4 mV/m, for the actual field strength to give a picture which would satisfy 70% of observers. The local variation factor, however, has also to be taken into account in arriving at the median field strength which is necessary to give a 70% probability of service. The appropriate calculations, based on field-strength measurements, led to the conclusion that, for the type of country in the Greater London area, the average value of required median field was about 3 mV/m. This value may be compared with those quoted in the last column of Table 5, with which it is in good general agreement.

Turning to Band I, although the figure of 0·1–0·15 mV/m stated to be necessary at an individual receiving-aerial location

corresponds well with recognized procedure, whereby 100 μV/m median field determines the boundary of the service area of a Band I television station, such a field is often found to be insufficient in practice. In the experiments which have been quoted it was found, for example, that as much as 1 mV/m ambient field was necessary to give a picture as good as those which had been used as the criterion for the Band V transmissions. The requirement for this higher field strength is undoubtedly due to the effect of impulsive interference and of 'pattern' interference on the picture. The latter type of interference from distant sources was much in evidence, owing to tropospheric and ionospheric activity at the time of observation.

Mr. S. N. Doherty: With regard to the allowance in Band I of 2 dB for cosmic noise (Table 2), I find from several references and by measurement that there is considerable doubt as to its precise value, but the general opinion is that up to 6 dB should be allowed on a receiver of 6 dB noise factor at 55 Mc/s. This effect is masked in urban areas by ignition interference.

The United States has recently completed an extensive survey of v.h.f. and u.h.f. service areas. Some conclusions are given in the following extract:*

> Death knell to u.h.f. television was sounded . . . in a report by the Television Allocations Study Organisation after a 2½ year investigation. . . . The report is expected to spur F.C.C. efforts to swop some u.h.f. spectrum space with the military to make room for a few more v.h.f. channels in a continuous band. Though u.h.f. is 'as good as' v.h.f. in short range over level terrain and is less subject to interference, these (following) deficiencies weighed heavily: the u.h.f. signal deteriorates rapidly as distance increases; the antenna and receiver are less efficient than v.h.f. and a u.h.f. station is more expensive to operate.

The military authorities are discussing the exchange with the F.C.C.† It might appear the United States have built their u.h.f. system first and tested it afterwards. In this country we have done it the other way round by thoroughly testing a typical service area first.

The opportunity still exists in this country and Europe to benefit by our own and extensive American u.h.f. experience, and negotiate for the 216–450 Mc/s band for television instead of the u.h.f. band. This would result in larger service areas per transmitter and less dissatisfied viewers within the service areas, owing to improved local variation factors compared with u.h.f. service.

Mr. J. K. S. Jowett: Of the various factors into which the author has broken down this problem, I think the two most controversial are the signal/noise ratio required to give a satisfactory picture and the percentage of viewers which one should satisfy in order to define the edge of a service area.

The author comments, rightly, on the 'subjectiveness' of any assessment of the first of these two factors. I think that the wide spread of signal/noise ratios which may prove acceptable in a variety of different situations cannot be over-emphasized. But for planning we want, if we can, to take a single representative figure and must not be swayed too much by what may be acceptable in extreme situations of one form or another. On this basis, I think that the author's choice of 12 dB is a fair one.

As to the percentage of satisfied viewers which may be taken to define the edge of a service area, I think this should be rather greater than the 50% which has been accepted in the past. But does the author not think it time that all service-area maps were produced to show simply, or primarily, contours of given percentage rates of satisfaction, rather than contours of median field strength? This subject has been ventilated already in discussions of the C.C.I.R., but I would hope that broadcasting organizations will go further than this and actually begin to produce maps for the public use showing contours of service grades defined in this new way.

In his Conclusions the author refers briefly to the special situation which may arise when considering the level of occasional fringe-area interference between co-channel stations. This is because the statistics he has introduced for local variation factor refer to the variations with local siting of the wanted signal only. Account has to be taken also of the local variations of the unwanted interfering signal. In one extreme case, when both wanted and unwanted signals come from the same direction, these two sets of local variations may be closely correlated, and this is the easiest condition to meet as regards planning for minimum interference. At the other extreme, however, the two signals may come from opposite directions and there may be an appreciable degree of negative correlation between their local variations of amplitude. In such cases, which may, for example, occur on the south-east coast of England when considering interference from Continental stations, it may be desirable to protect wanted median field strengths down to a lower-than-normal level in order to ensure satisfactory protection of a town or area as a whole. This subject does, in fact, require further investigation.

The subject of the paper is thus one of some complexity as well as importance. I think that the author has made an admirable start in trying to break it down into the various significant factors, although the precise values to be attached to the various factors are bound to be the subject of much further consideration and debate.

Mr. E. M. Lee: One ought to say more about people who are not within the 70% satisfied viewers. The 30% in Bands I and III are not too badly off; they are getting a slightly deteriorated picture, only a few decibels down; but in Bands IV and V there are many who do not get a picture at all.

The B.B.C.'s tests in Bands IV and V seemed to be not too discouraging after about five months' run, whereas in the United States, after running for a full year and spending fabulous sums trying to improve viewers' aerials, the experience has been so disastrous that they are having to shut transmitters down. The advertising sponsors are causing this shutting down, and we can imagine a similar situation arising here, with an advertiser objecting to paying for an appearance in Band V if his competitor has an appearance in Band III with a coverage of some 95% of the viewers in his area.

I was with N.B.C. engineers a month ago who had had the job of shutting down the stations after having spent a year trying to satisfy their sponsors by improving the viewers' aerials around Buffalo and similar places. We cannot be confident that 70% satisfied viewers will do; we want to know how dissatisfied the other 30% are.

THE AUTHOR'S REPLY TO THE ABOVE DISCUSSION

Mr. G. F. Swann (*in reply*): Mr. Jackson has produced some very useful information on the effect of ambient lighting on the signal/noise ratio for tolerable reception conditions, and I am glad he was able to confirm that a 12 dB ratio is appropriate to modern viewing conditions.

In reply to Mr. Anderson, it would, of course, be desirable to specify the signal/noise ratio in terms of the percentage of observers satisfied, but this might call for a rather too rigid

* *Electronics*: 27th March, 1959, p. 11.

† Doerfer, J. C.: 'F.C.C. Recommendations to Congress on Allocations', *Television Digest* (special supplement), 27th April, 1959.

specification of ambient lighting and other viewing conditions for general use in the present context. I feel we can do no more than adopt a 12 dB ratio as a generally tolerable grade of service when all relevant factors, which can be summed up as 'modern viewing conditions' are taken into account.

With regard to the noise contributions of video tape recordings and distribution links, the type of noise they produce is important. Television cameras and microwave links tend to have a 'triangular' noise distribution over the video band, and their noise is less disturbing than thermal noise of the same r.m.s. level. I would be surprised if any general allowance for programme-source noise was justified in arriving at the nominal service-area limit of a station, although this noise may, as he has found, make an appreciable contribution to the overall noise level under extreme conditions.

Mr. Townsend's emphasis on the public-relations problems which the large local u.h.f. variations can present is very much to the point, as also is Mr. Lee's concern for the 30% of locations at the 70% probability contour who will receive less than the required field strength. In average urban conditions and with random aerial siting, perhaps 1% of viewers along this contour would receive a signal/noise ratio of 6 dB, i.e. below the 'satisfactory' level, in Band I. For u.h.f. signals this percentage might be around 20%. Careful optimization of aerials would give considerable improvement, but would leave a residue of a few per cent who would not receive a satisfactory service. Mr. Townsend raises some pertinent questions in connection with possible higher-definition and colour services for which answers will need to be found, but which take us outside the admittedly limited scope of the paper. In reply to his doubts about the commercial practicability of an 8 dB noise factor for u.h.f. signals, I agree that improved valves will be needed before this can be achieved; whether these will be forthcoming, or whether current developments in the field of parametric amplifiers will provide an alternative solution to this problem, must remain to be seen. I am relieved to hear from Mr. Howe that the large-scale experiment to which he refers generally confirms the conclusions of the paper in respect of the field strengths required in the absence of man-made interference. The possibility of limitations to reception more serious than those due to nature and equipment performance is very well illustrated by the Band I results which he describes, and I am grateful to him for emphasizing that the practical situation, because of interference, may be different from that predicted by a consideration of receiver noise alone.

In reply to Mr. Doherty, I have found difficulty, as he has, in reconciling the increase in noise level due to cosmic noise observed in television installations with the generally higher values predicted by published cosmic-noise levels. The allowance of 2 dB made in Table 2 was based on measured increases at quiet sites, and further observations made since the paper was written suggest that a rather higher allowance of up to 4 dB would be more appropriate. His comments on recent American appraisements of the u.h.f. position are appreciated, and while I agree with him that we should make full use of American u.h.f. experience, we must also avoid giving undue weight to commercial aspects of u.h.f. service as they apply in America.

I share with Mr. Jowett a desire to see some form of television coverage presentation come into use which gives an indication of the probability of satisfactory service. The median field-strength contour map which is used at present, and which, if detailed, is so useful for engineering purposes, could be modified by showing percentages of 'satisfactory' locations in the towns to emphasize the effects of local variation. In view of the characteristics of a u.h.f. coverage, I hope that some such presentation will be used when the service areas of a possible u.h.f. service need to to be illustrated.

While only a very brief mention of the existence of channel-sharing problems was made in the paper, their importance was underlined by Mr. Howe's remarks, and Mr. Jowett's comments indicate the complexity involved when the local variation of both wanted and unwanted fields is considered. Although the probability that a location will receive adequate protection from a single source of interference is readily calculable if the variations of the two fields are assumed to be uncorrelated, this is clearly an insufficient approach for the type of situation he described. Further complications arise when several sources of interference exist, and the subject does, as he suggests, merit further investigation.

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FURTHER STUDIES OF THE DEVIATION OF LOW- AND MEDIUM-FREQUENCY GROUND WAVES AT A COAST-LINE

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(*The paper was first received* 26*th November*, 1958, *and in revised form* 16*th April*, 1959.)

SUMMARY

The paper opens with a review of the theoretical studies of coastal deviation which have been made in recent years. It shows that there is close agreement between the various formulae derived for the deviation and that the graphical method used in an earlier paper by two of the present authors produces results which are substantially in agreement with those obtained from the formulae.

In an endeavour to obtain experimental confirmation of the theoretical magnitude of the deviation and its variation with distance from the boundary with angle of incidence of the wave front at the boundary, etc., existing data from various sources have been examined. Two new series of measurement on low- and medium-frequency waves have also been made. In both cases the transmitters were located at sea and the directional measurements taken at numerous sites on land. The general conclusion obtained from this experimental investigation was that a deviation due solely to the change in ground conductivity at the boundary and of the magnitude given by theory is small compared with the larger random directional errors which are attributed to ground irregularities, surface obstructions and other causes. The only systematic deviations observed had amplitudes of up to 7° and occurred within a few hundred metres (less than a half-wavelength) of the coast on the seaward side; at such near distances the theoretical formulae are not valid. It is concluded that for many coastal direction-finding stations coastal deviation as such is unimportant.

(1) INTRODUCTION

The difference between the electrical constants of the sea and those of the land leads to a deviation in the direction of travel of ground waves propagated across a coast-line. In a previous paper[1] it was shown how the deviation could be calculated from the change of phase of the wave with distance along the path of propagation. Furthermore, it was pointed out that this method took into account the phase disturbance which was found to occur near a boundary and that such a disturbance should give rise to a rapid increase in the deviation as the receiving point approached the coast-line. This deduction was based on the assumption that the coast-line was straight and well defined and that the land surface was smooth and homogeneous.

Application of this theory to existing experimental data showed that the increase of deviation near the coast might provide the explanation of the fact that observed deviations were considerably greater than those previously calculated by a simple method which neglected the local phase disturbances. However, in an investigation[1,2] of the phase change with distance of l.f. waves propagated across a coast-line and of the associated deviations of the phase front along the path it was found impracticable to obtain any definite confirmation of the expected variation of deviation with distance. This was due to the presence of other phase disturbances, which were attributed to irregularities in the coast-line or undulations of the land, and resulted in variations of the deviation which masked the wanted effects. The coastal regions involved were not ideally suited to this type of directional investigation; they had been selected primarily for the phase-change measurements. It was considered useful, therefore, to carry out further directional measurements at a more suitable site and on a higher frequency. A few measurements, however, were first made over land at the original site. Both series of investigations are described in the paper.

Since the first paper was written a number of papers have been published on the theory of propagation over a boundary, and formulae for the deviation are given in them. These theories are reviewed in the present paper and comparisons are made with the experimental results.

(2) THEORETICAL

(2.1) Review of Published Work

The first satisfactory theoretical discussion of mixed-path ground-wave propagation and coastal deviation was given by Grunberg.[3] He considered a plane wave propagated across a sharp discontinuity in the electrical properties of a flat earth, and assumed one of the earth media to be perfectly conducting. He showed that the deviation decreased with increasing distance from the coast.

Later Feinberg[4] generalized Grunberg's treatment by taking the transmitter to be a vertical dipole at a finite distance from the boundary. He defined the deviation angle, δ, as that formed by the intersections with the ground of the disturbed and undisturbed surfaces of constant phase. Writing the phase of the field in the form $\phi' = kr + \phi$, where $k = 2\pi/\lambda$ is the free-space wave number and ϕ is the phase lag, he showed that the deviation is given by

$$\tan \delta = \frac{-\frac{1}{r}\frac{\partial \phi}{\partial \theta}}{k + \partial\phi/\partial r} \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where (r, θ) are the polar co-ordinates of the receiver, R, referred to the transmitter, T, as shown in Fig. 1(*a*), and θ and δ are measured in a clockwise direction. When $k \gg \partial\phi/\partial r$, which is so in practice except perhaps when the receiver is very near the boundary, we have, for small values of δ,

$$\delta = \frac{-1}{kr}\frac{\partial \phi}{\partial \theta} \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

Using an approximate boundary condition, namely that the ground exhibits surface impedance, Feinberg derived an integral equation for the field, which he solved for various positions of the transmitter and receiver by making approximations so as to obtain known functions in the integral. Then using eqn. (2) he obtained expressions for the deviation angles.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

The paper is an official communication from the Radio Research Station, Department of Scientific and Industrial Research.

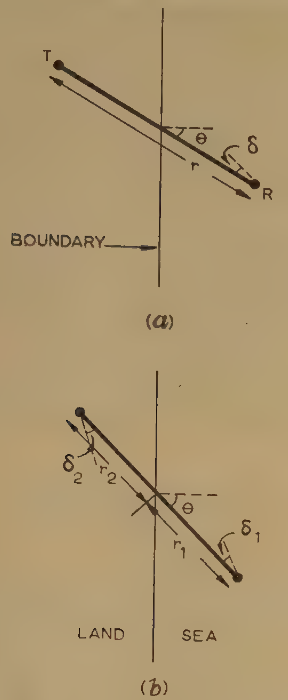

Fig. 1.—Path configurations.

(a) Corresponding to eqn. (1).
(b) Corresponding to eqn. (3).

Two of these, valid when $kr_1 \cos^2\theta \gg 1$ and $kr_2 \cos^2\theta \gg 1$, and when the numerical distances* are not large, are

$$\delta_1 = 1\cdot18 \times 10^{-7}\sqrt{\left(\frac{r_2}{r_1}\right)}\frac{\tan\theta}{\sqrt{(\sigma r)}} \quad . \quad . \quad (3a)$$

$$\delta_2 = -\,1\cdot18 \times 10^{-7}\sqrt{\left(\frac{r_1}{r_2}\right)}\frac{\tan\theta}{\sqrt{(\sigma r)}} \quad . \quad . \quad (3b)$$

where r_1 and r_2 = Lengths of the paths over sea and land, respectively ($r_1 + r_2 = r$), km.

δ_1 and δ_2 = Deviations observed on sea and land, respectively, positive angles being measured in a clockwise direction as shown, deg.

σ = Conductivity of the land [see Fig. 1(b)], e.m.u.

It is of interest to note that the deviation given by eqn. (3) is independent of frequency.

In a more recent paper Furutsu[5] considered a vertical dipole transmitter above a flat earth comprised of two homogeneous finitely conducting sections separated by a sharp boundary. For the case when the transmitter and receiver are near the ground, he was able to derive an expression for the Hertzian vector, which is valid when the distances of the transmitter and receiver from the boundary are such that $kr_1 \cos\theta$ and $kr_2 \cos\theta$ are large. When the transmitter and receiver are actually on the ground and one of the media is perfectly conducting, the Hertzian vector reduces to

$$\Pi = 2A\Pi_0 = 2\Pi_0\left\{K(\rho) + \frac{4j}{\sqrt{\pi}}\varepsilon^{-\rho}\sqrt{\rho}\,U\left[j\sqrt{\rho_1}, \sqrt{\left(\frac{r_2}{r_1}\right)}\right]\right\} \quad (4)$$

where Π_0 is the free-space field,

$K(\rho) = 1 - 2j(\sqrt{\rho})\varepsilon^{-\rho}\int_{j\sqrt{\rho}}^{\infty}\varepsilon^{-x^2}dx$, is the Sommerfeld attenuation function,

ρ, ρ_1 and ρ_2 are the numerical distances corresponding to the distances r, r_1 and r_2 respectively,

$$U(\alpha, n) = \int_{\alpha}^{\infty}\int_{nx}^{\infty}\varepsilon^{-x^2}\varepsilon^{-y^2}dxdy$$

and a time factor $\varepsilon^{+j\omega t}$ is assumed.

Furutsu defined the deviation angle to be the angle between the normal to the horizontal magnetic field and the line joining the transmitter and receiver. He showed that the magnetic field is elliptically polarized, but that the observed deviation angle is given to the first order by

$$\delta = \mathscr{R}\left(\frac{j}{kr}\frac{1}{A}\frac{\partial A}{\partial\theta}\right) = -\frac{1}{kr}\frac{\partial\phi}{\partial\theta} \quad . \quad . \quad . \quad . \quad (5)$$

This is identical with eqn. (2) derived by Feinberg. Substituting the value of A given by eqn. (4) into eqn. (5) gives[5]

$$\delta_1 = \frac{1}{\sqrt{\pi}}\sqrt{\left(\frac{r_2}{r_1}\right)}\frac{\tan\theta}{kr}\mathscr{R}\left[\sqrt{\rho}\frac{K(\rho_2)}{A}\right] \quad . \quad . \quad (6a)$$

$$\delta_2 = \frac{-1}{\sqrt{\pi}}\sqrt{\left(\frac{r_1}{r_2}\right)}\frac{\tan\theta}{kr}\mathscr{R}\left[\sqrt{\rho}\frac{K(\rho_2)}{A}\right] . \quad . \quad . \quad (6b)$$

When the numerical distance of the land path is small, i.e. when $|\rho_2| \ll 1$, $K(\rho_2)/A \simeq 1$, and if displacement currents are neglected, eqn. (6) reduces to eqn. (3), in agreement with Feinberg.

Wait[6] considered the same earth model as Furutsu, but took the transmitter and receiver to be on the ground. He derived an integral expression for the field which is in agreement with the previous results, and plotted curves of the amplitude and phase for a wide range of parameters. He also obtained an expression for the deviation which can be shown to be equivalent to eqn. (2). This expression together with his field curves can be used to find numerical values of the deviation by a graphical method when both earth media are finitely conducting (or one perfectly conducting).

Senior[7] has derived simple expressions for the deviation for various positions of the transmitter and receiver when the sea is assumed to be perfectly conducting and the land a perfect dielectric. He has also given an expression which is valid when land is conducting, the displacement current being negligible, and when the transmitter is far out over the sea. When the numerical distance over land is much less than unity, this reduces to the expression given by eqn. (3b).

The validity of all the results discussed above seems to depend on the factors $kr_1 \cos\theta$ and $kr_2 \cos\theta$ being large. This means, in effect, that the transmitter and receiver are at such distances from the boundary that only the radiation component of the field need be considered.

The practical conclusions to be drawn from this review are that, at low and medium frequencies and subject to the above limitation, the deviation can always be obtained either from eqn. (6) or by the graphical method suggested by Wait. However, if the numerical distance over land is much less than unity eqn. (3) will give the deviation to a first order of accuracy.

(2.2) Comparison with Previous Graphical Method

In the previous paper[1] it was shown how the deviation could be derived graphically from curves of phase change with distance. These phase-change curves for the inhomogeneous path were obtained by a method[8] analogous to that introduced by Millington[9] for the determination of amplitude over a mixed

* The numerical distance is a function of path length, r, frequency, f, and ground constants; for low and medium frequencies and for all but very low ground conductivities it may be written as $\rho = 5\cdot6\pi rf/\lambda\sigma \times 10^{-22}$.

path. The deviation formula obtained for the case of the transmitter at a large distance from the coast-line was

$$\tan\delta = \frac{\lambda}{2\pi}\tan\theta\left[\left(\frac{d\phi}{dr}\right)_c - \left(\frac{d\phi}{dr}\right)_x\right] \quad . \quad . \quad (7)$$

where $(d\phi/dr)_c$ is the slope of the phase curve at the boundary on the transmitter side and $(d\phi/dr)_x$ is the slope at the point at which the deviation is being determined.

This formula may be generalized so as to apply to cases where the transmitter is at a finite distance from the boundary by introducing a distance factor to give

$$\tan\delta = \frac{d}{D+d}\frac{\lambda}{2\pi}\tan\theta\left[\left(\frac{d\phi}{dr}\right)_c - \left(\frac{d\phi}{dr}\right)_x\right] \quad . \quad . \quad (8)$$

where d = Distance of transmitter from boundary.
D = Distance of receiving point from boundary, measured along the path of propagation.

It is not practicable to compare eqn. (8) with eqn. (3) or (6) analytically. The results of using these expressions in particular cases can, however, form a basis of comparison and Fig. 2 shows four such examples. It is apparent that, under practical ground conditions and for low and medium frequencies, the deviations computed from eqns. (3) and (6) show good agreement with those obtained graphically using eqn. (8), particularly at the medium frequencies.

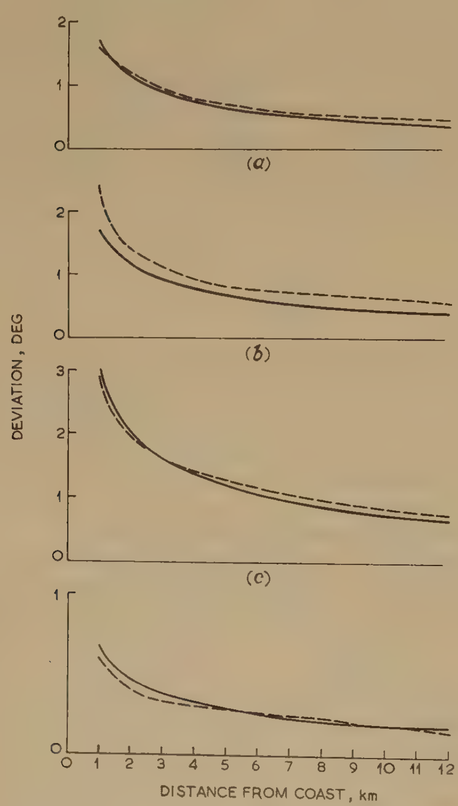

Fig. 2.—Comparison between methods of deriving deviation.

———— Computed from eqn. {(6) in (a) and (c); (3) in (b) and (d).
– – – – Determined graphically using eqn. (8).
(a) f = 500 kc/s $\sigma = 3 \times 10^{-14}$ e.m.u. $\theta = 70°$
Transmitter on land 20 km from coast.
(b) f = 127·5 kc/s $\sigma = 3 \times 10^{-14}$ e.m.u. $\theta = 70°$
Transmitter on land 20 km from coast.
(c) f = 300 kc/s $\sigma = 10^{-14}$ e.m.u. $\theta = -70°$
Transmitter at sea 40 km from coast.
(d) f = 300 kc/s $\sigma = 2 \times 10^{-13}$ e.m.u. $\theta = -70°$
Transmitter at sea 30 km from coast.

(3) EXPERIMENTAL EVIDENCE

Reliable experimental evidence of the existence of coastal deviation is scarce. Two sources, that of Eckersley's observations in Cyprus[10] and that of Smith-Rose's observations at Orfordness,[11] both on medium frequencies, were discussed in the first paper. Both provide rather limited data.

Another source of data is the calibration of coastal direction-finding stations, and the results of an examination of a group of these are given below.

In an endeavour to obtain fresh evidence two series of measurements were made on medium frequencies using portable direction-finders; one on the south coast in the same area as that used for the l.f. measurements[1] and the other on the east coast of Yorkshire. In both cases the transmission was across the boundary from the seaward side; this is the converse of the conditions under which the low-frequency measurements were conducted.

(3.1) Bearing Errors on Established Direction-Finding Stations

The errors on a number of land-based direction-finding stations situated near the coast were examined for evidence of coastal deviation. The data used consisted of the bearing calibrations of ten selected coastal stations which were sited in such a way that some coastal-deviation effects might be expected. The calibrations had been made with a shipborne transmitter at a radius of several miles on a frequency of 500 kc/s and on a second frequency of either 375 or 410 kc/s.

An analysis of the error curves showed the existence of instrumental errors containing semicircular and quadrantal components up to 3° in amplitude and octantal components up to 4°. There were also random errors having peaks up to 4°.

Careful examination of the error curves over those sections in which the path of propagation lay nearly tangential to the coast produced no evidence of any deviations which might be attributed to the boundary. It is to be noted, however, that the magnitudes of the deviations which, according to theory, are to be expected at an angle of incidence to the coast of 10° and under the conditions applicable to these stations would not exceed 1·5°. So that, even if errors of this magnitude existed, they would not be readily detected in the presence of the larger instrumental errors. The only marked bearing changes observed which might be attributed to coastal effects occurred on those azimuths on which the ship carrying the calibrating transmitter might have been passing, or emerging from, behind a headland.

The general conclusion from this examination of data from coastal direction-finding stations is that any errors due to coastal deviations which could be expected to be present are masked by instrumental errors and by random errors due to ground configuration or other causes.

(3.2) South-Coast Measurements

The south-coast series of directional measurements were made behind the straight stretch of coast-line between Peacehaven and Rottingdean, Sussex. The cliffs there are about 30 m high and the hinterland is open chalk downland having a conductivity of about 10^{-13} e.m.u. The transmissions received were from five marine beacons, which were located on the French coast, in the Channel Islands and at Start Point and operated on frequencies in the range 290–310 kc/s, and also from an aeronautical beacon at Dieppe on a frequency of 350 kc/s. The azimuths of the transmitters ranged from +58 to −62°, with respect to the normal to the coast-line.

The bearings were taken on a rotating-loop direction-finder mounted on the roof of a small van. The instrumental errors

due to the van were of the quadrantal type and had a peak value of about 1°; all observations were corrected accordingly. The bearing scale was oriented to an accuracy of $\frac{1}{4}$° by visual observations on known landmarks. The observations were made at 19 sites lying on two straight lines of which one was normal to the coast and the other was at an angle of −60° to the normal. All the sites were free of obstructions and would be expected to have negligible site errors.

The observed bearing errors were plotted against the distance of the site from the coast measured along the line of propagation. This was done for each transmitter and both lines of sites. Three examples of these plots are shown in Fig. 3.

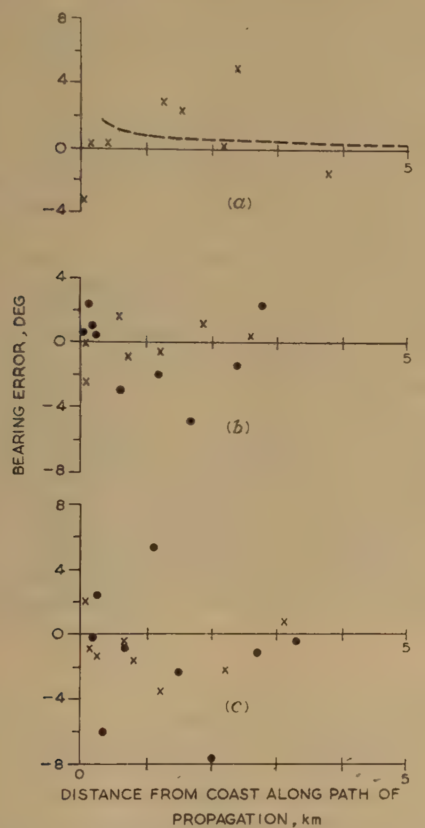

Fig. 3.—Directional errors on the south coast.
- - - - Theoretical bearing error for $\theta = -62°$.
● Observed points along line normal to coast.
× Observed points along line at −60° to normal.

(a) Dieppe beacon. $\theta = -62°$
(b) Barfleur beacon. $\theta = 22°$
(c) Casquets beacon. $\theta = 36°$

The main deduction to be made from these measurements is that in all cases there is a large random component of error which is as great when the angle of incidence is near the normal to the coast [Figs. 3(b) and (c)] as it is when the incidence of the wave is more oblique [Fig. 3(a)]. It is also to be noted that there is a marked difference between the two series of measurements on the Barfleur beacon and those on the Casquets beacon. These differences and the general randomness of the errors indicates that they are determined largely by the sites used for the measurements rather than by the distance from the coast, the angle of incidence of the waves or the ground characteristics. The only systematic feature of the errors which is detectable is that the values are generally positive when the angle of incidence is negative [Fig. 3(a)] and negative when the angle is positive [Fig. 3(c)]. This is in accordance with the theory.

The deviation calculated from eqn. (3) for an angle of incidence of −62° is shown in Fig. 3(a). It is seen that except for points very near the coast, the values are considerably smaller than the observed errors, and it is clear that if deviations of this order were actually present they would be masked by the random errors.

(3.3) Yorkshire-Coast Measurements

A series of measurements was made in the coastal region lying south of Bridlington and north of the Humber. Observations were made on three transmitters, the principal one being the marine beacon situated on Flamborough Head and operating on a frequency of 303 kc/s. The second transmitter was the Spurn-Light-Vessel beacon operating on the same frequency (but on a time-sharing basis) and anchored about 10 km off Spurn Head. The third was the Droitwich transmitter (200 kc/s) which was just over 200 km distant in a south-westerly direction. The positions of these transmitters relative to the coast-line and area of operations are shown in Fig. 4. Also in this Figure

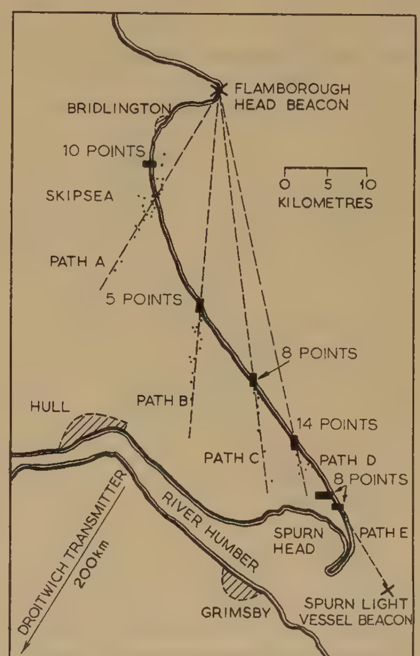

Fig. 4.—Sites of east-coast measurements.

are shown the approximate positions of the sites at which directional measurements were made, and the directions of the paths of propagation appropriate to the five major groups of sites are indicated. It will be seen that from Flamborough Head angles of incidence to the coast ranging from 45 to 70° can be obtained while the paths from the Spurn Light Vessel cross the coast at nearly grazing incidence.

The coastal area covered by the measurement sites was generally flat with no contour variations greater than about 15 m and with no cliffs higher than 20 m. It was exceptionally free from wooded areas, towns and obstructions which might cause errors. The estimated average ground conductivity was 2×10^{-13} e.m.u. Compared with the south-coast region this area was appreciably flatter with lower cliffs but with a higher conductivity.

The measurements were again made with a rotating-loop direction-finder which was placed on the ground and not mounted on a van roof as before. This arrangement not only allowed

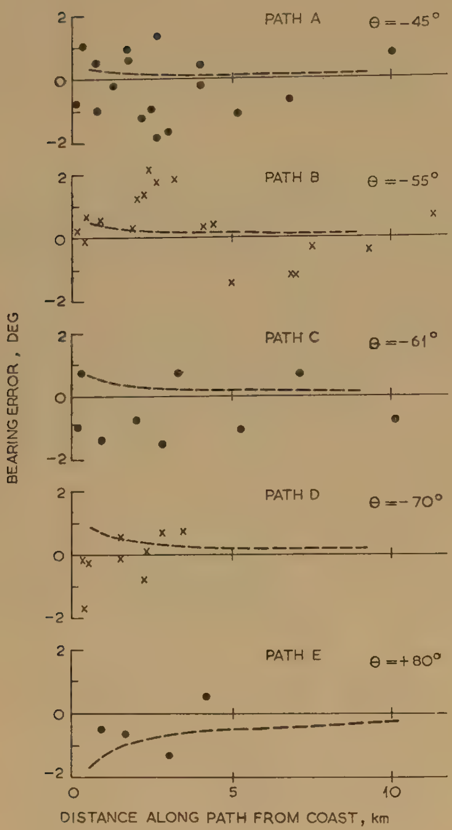

Fig. 5.—Directional errors on east coast.
× ● Measured errors.
– – – – – Theoretical errors.

greater freedom in the choice of site but also eliminated the need to correct for errors due to the van. Observations on one or more of the transmitters were made at all the sites shown by the points in Fig. 4. Small areas containing a large number of sites are indicated by a single block marked with the number of points. It will be noted that a number of measurements were taken at sea. For these the direction-finder was installed in a small wooden motor launch whose position was fixed by sextant observations on known landmarks. Since the transmitter site (Flamborough Head) was visible from the launch it was possible to observe the bearing error directly using the special optical system already described.[1]

In Fig. 5 are shown the errors observed on land plotted against distance from the coast for the five paths. They show a wide scatter of the readings with no definite trend with distance or with angle of incidence. Nor do they even show any definite change in the sign of the error as the angle of incidence changes from negative to positive as did the results of the south-coast measurements (Fig. 3). The broken lines show the theoretical deviation calculated from eqn. (3) and it is seen that the changes of observed error from one site to the next is many times greater than this deviation.

A mass plot of all the measurements obtained within one wavelength of the coast on land, on the sea-shore and at sea is shown in Fig. 6. They are plotted according to their distance along the path of propagation relative to the cliff edge In contrast to the results shown in Fig. 5 these errors measured on the coast-line (Fig. 6) show a systematic trend taking the form of an error peak on the seaward side of the cliff which reverses in sign when the angle of incidence passes through zero (compare errors of group 6 with those of the other groups). The distance between the cliff and the water's edge varied from one group of sites to another. In the case of groups 3 and 4 the distance was approximately 100 m but for group 2 it was about three times as great although the angles of incidence were the same. If this is taken into account, together with the fact that

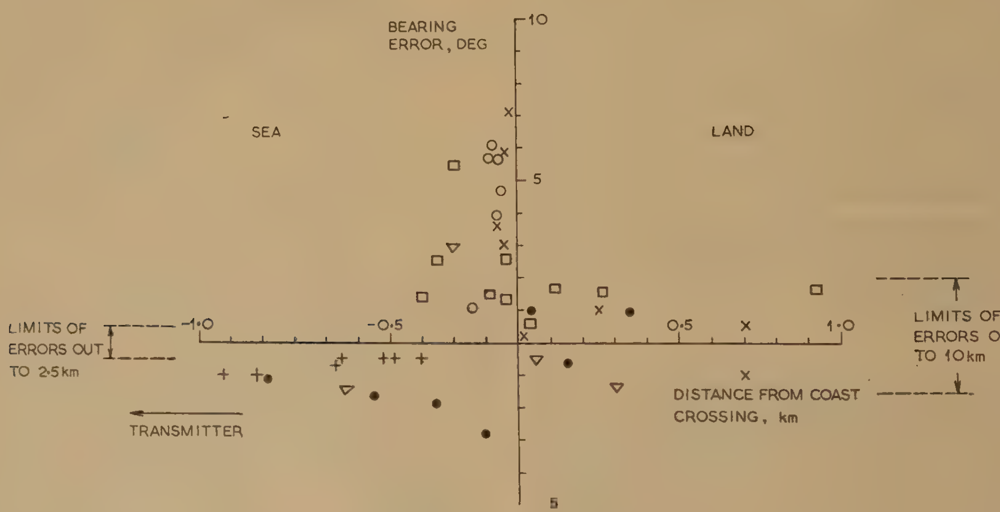

Fig. 6.—Directional errors at coast.

Group	Area of observation	θ	Transmitter
		deg	
1	+ At sea	−45	Flamborough
2	□ Near Skipsea	−45	Flamborough
3	○ Near Skipsea	−45	Flamborough
4	× Near Skipsea	−45	Flamborough
5	▽ Near Spurn Head	−83	Flamborough
6	● Near Spurn Head	76	Spurn Light Vessel

also in the case of groups 5 and 6 the distances along the paths to the water's edge were greater because of the larger angles of incidence, it will be found that most of the rise in deviation occurs over the sands which at their widest did not exceed 300 m, i.e. 0·3 wavelength.

An example of the measurements made on the Droitwich transmitter is given in Fig. 7. This shows the variation in

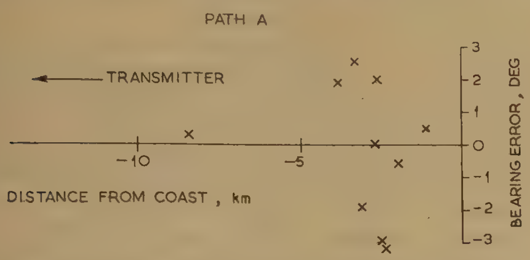

Fig. 7.—Directional errors on Droitwich transmitter.

bearing along a line extending inland from the coast towards the transmitter (path A) and is therefore indicative of the variation in the area due to field disturbances other than those resulting from transmission across the coastal boundary. Comparison with the observations made in the same area on the Flamborough Head beacon (Fig. 5) shows that the nature and magnitude of the variations are similar in both cases and supports the conclusion that the coastal boundary is not the major cause of the observed errors. Further support has also been provided by measurements on Droitwich made near Slough in a large open area where, over a distance of 1·5 km, bearing-error variations up to 3·5° were observed. In contrast it should be noted that the scatter of the errors observed at sea (more than 1 km from the coast) on the Flamborough Head transmitter (Fig. 6) were appreciably smaller than this—in fact, the scatter of $\pm\frac{1}{2}°$ indicated in the Figure is of the order of the accuracy of measurement at sea.

(4) CONCLUSIONS

The examination of the theoretical work on the subject has shown that a large measure of agreement exists between the various formulae obtained for the deviations at a boundary. The formula given in the previous paper, which was derived from consideration of phase/distance curves, is not identical with those based on more fundamental considerations, but the difference appears to be of the second order and for most practical purposes can be regarded as equivalent to them. The latter formulae, however, have been used in the present paper to calculate the theoretical deviations, the simpler expression [eqn. (3)] being employed whenever possible in preference to the more complex one [eqn. (6)].

The main features of these expressions [eqns. (3) and (6)] are:

(*a*) To a first approximation, the deviation is independent of frequency.

(*b*) The deviation increases as the receiving point approaches the coast and decreases as the transmitter approaches.

(*c*) The deviation is proportional to the tangent of the angle of incidence of the wave.

The distinction between transmitter and receiver movement (*b*) follows from a consideration of the geometry of the ray paths and applies to transmission both from land to sea and vice versa. It is independent of any local phase distortions at the coast but the magnitude of the deviation changes will be modified by these phase distortions in such a way as to exaggerate the changes for receiver movement and diminish those for transmitter movement.* It is to be noted also that the deviation is proportional to the tangent of the angle of incidence only if the lengths of the land and sea paths remain constant as the angle changes. In the practical case where a direction-finder is in a fixed position and the distance of the transmitter from the coast is large compared with that of the receiver, the deviation is proportional to $\tan\theta(\cos\theta)^{1/2}$. The expressions are not valid for all values of their parameters but, generally, they apply in the low- and medium-frequency ranges and when the transmitter and receiver are more than about a half-wavelength from the coast.

The calculated errors for most practical conditions are very small, being less than 1°, except perhaps at positions within a wavelength of the boundary, or where the land conductivity is very low or when the wave crosses the coast at grazing incidence.

Attempts to measure the deviation due solely to the change in ground conductivity have been thwarted by the presence of larger errors of a random nature which are attributed to ground irregularities and other causes not associated with the boundary. To find a site on which such disturbances would be small compared with the coastal deviations appears to be very difficult since the requirement is for a long straight coast-line with negligible cliff and land of low uniform conductivity. Such ideal coastal conditions are uncommon since, although the coast line may be straight and cliffs negligible, as for the Yorkshire measurements, the ground conductivity is usually large because of the high water content of the soil in such low-lying areas. Where the ground conductivity is low, i.e. in a rocky region, there are usually cliffs, an irregular coast-line and an undulating hinterland containing areas of markedly different conductivity.

The only systematic deviation which has been shown by the measurements to exist is that which occurs on the coast itself, i.e. within about one-tenth of a wavelength of the effective boundary.

The general conclusion to be drawn from the theoretical studies and from the experimental work is that on sites of good or average conductivity coastal deviation as such is unimportant provided that the direction-finder is not sited too near the coast-line, i.e. within about 0·5 km for frequencies in the l.f. and m.f. bands. Although appreciable errors may still be observed these are not likely to be greater than those experienced on similar inland sites.[11] On a very-low-conductivity site substantial deviations are theoretically possible, but even here the nature of the coast and hinterland may often in practice be such as to give rise to errors of the same order as, or greater than, those due to the conductivity change at the coast. There may be exceptions, however, such as Eckersley's Cyprus site, but even in this case it seems possible that the errors may not have been primarily due to boundary effects.

The above conclusions will also apply to sea-borne direction-finders so far as the coastal deviation is concerned. But it must be remembered that at sea local-site errors, apart from those due to the ship, are absent and the influence of the land on the random errors may be evident at distances greater than 0·5 km out to sea. For example the earlier measurements[1] made at sea on 127·5 kc/s showed appreciable errors attributable to the land irregularities at distances of 5 km from the coast.

One is also led to the conclusion that the study of coastal deviation is mainly of academic interest and that, if confirmation of the results of the theoretical investigations is to be obtained experimentally, the model technique should be used. Although this method has its own peculiar difficulties, ideal boundary conditions can be attained and the order of the deviation due solely to the conductivity change determined.

* Throughout the paper the deviations considered are those of bearings determined from the directional characteristics of the receiving aerial. If, however, a directional transmitter and a non-directional receiver are used, then reciprocity indicates that the variations in the magnitude of the deviation with transmitter and receiver movements would be reversed, i.e. the deviation would increase as the transmitter approached the coast and decrease as the receiver approached.

(5) ACKNOWLEDGMENTS

The data on the calibrations of the coastal direction-finding stations referred to in Section 3.1 were provided by the Post Office Engineering Department. Mr. D. J. Barlow assisted in the observational work.

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A MULTIPLE-DIRECTION UNIVERSALLY-STEERABLE AERIAL SYSTEM FOR H.F. OPERATION

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(*The paper was first received* 17*th March and in revised form* 10*th June*, 1959.)

SUMMARY

The paper presents briefly the principles underlying the current development of a high-frequency directional aerial system which can be steered in both azimuth and elevation. The system comprises a number of omnidirectional unit aerials whose outputs (in the case of reception) may be brought approximately into phase with each other for any desired combination of frequency and direction by means of direction-control equipment which is basically a high-speed computer. The aerial system provides a number of independent outlets, each of which may be individually steered by the direction-control equipment. An experimental system based on these principles and designed for the reception of high-frequency signals from transmitters at long distances will be described in later papers.

The present paper mentions briefly some of the potential advantages of this type of aerial system, e.g. flexibility of design and operation, and notes that the basic principles are applicable to both transmission and reception.

(1) INTRODUCTION

The use of a highly-directive h.f. receiving aerial system capable of being steered in both azimuth and in elevation appears to offer the possibility of improved performance on h.f. radio circuits. On fixed point-to-point radio links such a system should be a valuable means of diminishing the effects of ionospheric movements normally experienced when fixed-direction aerials are used. An aerial array steerable in elevation has been in use on transatlantic radio circuits[1] for nearly 20 years: it provides an overall circuit performance which is superior to that obtainable with a single rhombic aerial, but its use is restricted to the New York–London route only. A universally-steerable aerial system could be used on any short-wave circuits and would be capable of coping, not only with variations of elevation angle[2] caused by changes in ionospheric conditions along the path of the signal, but also with any lateral deviations[3] arising from ionospheric tilts[4] or absorption in the auroral zones.[5] The system would provide a means of maintaining consistently high aerial gain when receiving signals from mobile stations on ships or aircraft. Moreover, it offers the possibility, in some circumstances, of reducing interference from unwanted transmissions.

In addition to the above uses for communication, such a system would have applications in direction-finding[6] and would also provide a powerful tool for propagation studies.

Proposals for an aerial system of this type were made by Bray,[7] but they were not readily realizable in a practical form. A more recent study has been made of the problem, and this has led to the development of a multiple-direction universally-steerable aerial system (medusa) which will provide all the above-mentioned facilities together with a high degree of flexibility to meet expanding operational requirements. The principles underlying the system are outlined briefly in the present paper. More comprehensive information will be made available in later papers, when the construction of an experimental aerial system for field trials has been completed.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Messrs. Morris and Mitchell are at the Post Office Research Station.

(2) PRINCIPLES

(2.1) General Considerations

The essential requirement for any h.f. directional receiving aerial system is that the voltages induced in the various elements of the system by the wanted signal shall be co-phased at the point where they are combined. So far as possible, these conditions should not apply for signals received from directions other than that of the wanted signal. With the usual type of fixed-direction array, e.g. rhombic aerial, horizontal array of dipoles (h.a.d.) or Koomans array, this requirement is met by the geometrical configuration of the array relative to the array axis, which is arranged to be along the azimuthal direction of the incoming wanted signal. Limited slewing facilities, applicable to both transmitting and receiving arrays, are sometimes provided by the insertion of phase-shift devices in, for example, a broadside array,[8] and arrangements for limited adjustment of the angle of elevation as well have also been described.[9] Recently a system based on the Wullenweber technique has been proposed in which full azimuthal coverage is provided.[10]

In the medusa system complete flexibility in azimuth and elevation is achieved by using a large number of substantially omnidirectional* aerials distributed over an extensive ground area: the phase of each aerial output is adjusted individually as required to co-phase all the outputs for the particular signal frequency and 3-dimensional direction desired. By the use of distribution units associated with the aerial outputs, the one aerial array can provide as many independent outlets as required; each outlet may be allocated for operation in a specific direction (i.e. for a particular radio link), but can be steered universally as necessitated by changing propagation conditions. A simplified schematic of the complete system is shown in Fig. 1 and a brief description is given in Sections 2.2–2.4.

(2.2) Aerial Array

The requirement of an omnidirectional aerial is met by an inverted-cone aerial designed for the reception of the vertically-polarized components of signals in the band 5–25 Mc/s. For this frequency band the response of the aerial is reasonably uniform over all azimuths and all angles of elevation up to 50°. (This range of elevation angle embraces all angles normally experienced on point-to-point and ship-to-land working, but would require modification for the reception of high-angle signals from aircraft or earth satellites.)

The array comprises a large number of these unit aerials distributed over an area having linear dimensions of several wavelengths. The aerial density is maintained reasonably uniform over the whole of the array to ensure representative sampling of the incident waves, but a regular array-pattern is avoided by using non-uniform spacings between individual aerials. This 'randomized' type of array is employed to reduce the possibility of appreciable side-lobes in the directivity charac-

* For complete directional coverage in azimuth and elevation the ideal unit aerial would be isotropic, but in practice it is sufficient if the response is reasonably uniform over the range of directions appropriate to the functional requirements of the system. Thus for long-distance point-to-point services, for example, the aerial response at elevation angles greater than, say, 50° is unimportant.

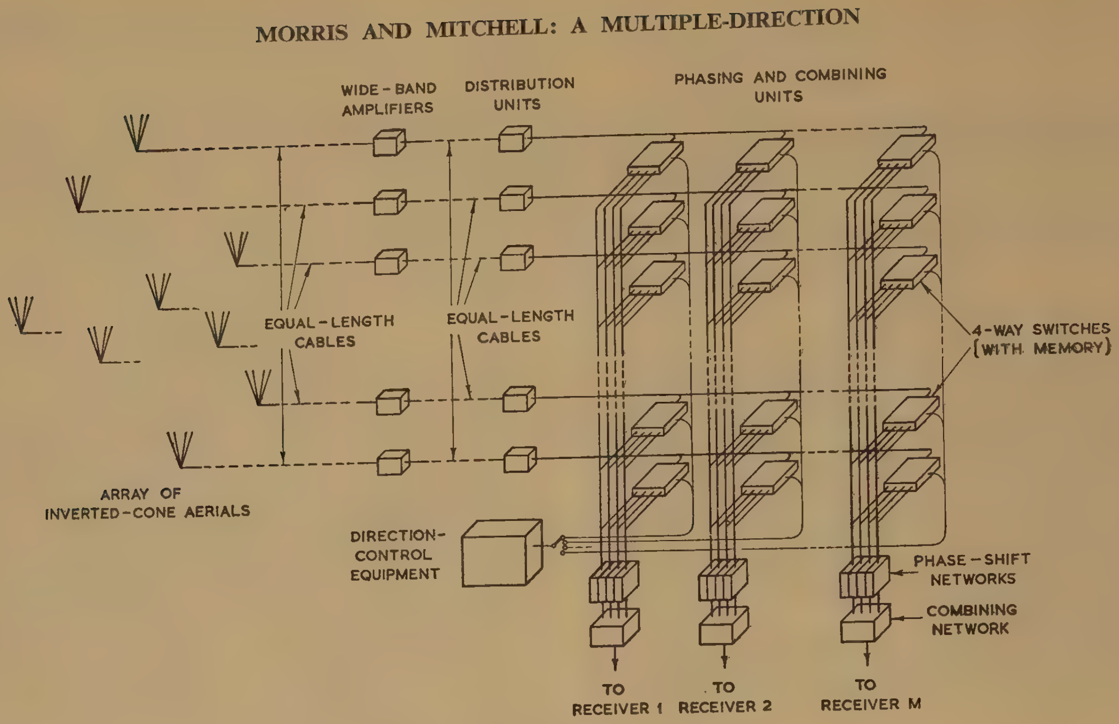

Fig. 1.—Schematic of aerial system.

teristics of the aerial system; this is discussed further in Section 2.3. The effects of mutual impedance between the aerials are rendered insignificant by spacing the aerials adequately.

(2.3) High-Frequency Equipment

Each aerial is connected by coaxial cables through a wide-band amplifier to a distribution unit which provides sufficient independent outlets to supply all the receiver inputs associated with the aerial system. Thus all the receivers can operate independently on different radio links, i.e. on different frequencies and directions. Considering any one receiver, wide-band signals are received from all the unit aerials in the system and are fed to a phasing-and-combining unit associated with the receiver. In this unit the wanted-signal contributions from all the aerials are brought substantially into phase with each other. The required values of relative phase-shift, which are quantized to avoid the provision of continuous phase-shifters, are determined for any desired frequency and 3-dimensional direction by centralized direction control equipment. This latter computes the appropriate phase shifts and then operates electronic switches (on the phasing-and-combining unit) to group together all the aerial contributions requiring the same phase shifts. Fig. 1 shows an arrangement based on four possible values of relative phase-shift, namely 0°, 90°, 180° and 270°. The four groups of signal contributions are passed through four corresponding wide-band phase-shift networks which introduce relative phase shifts of 0°, 90°, 180° and 270°, and thereby bring the four groups of outlets substantially into phase with each other (considering the wanted signal only). The outputs of the networks now include all the wanted-signal contributions of the individual aerials, all approximately in the same phase, and these are finally combined to form the 'steered' wanted signal input to the receiver.

For signals differing from the wanted signal in frequency and/or direction of arrival, the contributions from the various unit aerials will not, in general, be co-phased. Nevertheless, with an aerial array of regular pattern there would be certain combinations of aerials for which the phase shifts necessary to phase the wanted signal were identical with those for some other unwanted directions, resulting in appreciable side-lobes in the directivity characteristics of the aerial system. The 'randomized' array referred to in Section 2.2 has been adopted to reduce such side-lobes; this type of array ensures that, for each direction and/or frequency except that for which the system is steered, all the vectors representing the aerial contributions (after phasing the wanted signal) are randomly distributed in phase.

(2.4) Direction-Control Equipment

The direction-control equipment is a special-purpose digital computer for determining the individual 'phasing instructions' for each aerial in the system for any required frequency and direction of steer, specified in azimuth and elevation. Such equipment is inherently complex, but since it is capable of a high speed of working, one centralized equipment can control all the branches of the system—an arrangement which is consistent with current operational trends towards centralized control. Any one branch which requires steering has the sole use of the control equipment for the brief period necessary to determine the appropriate phasing instructions; during this period the other branches remain as last steered, under the control of memory devices associated with the switches.

Full facilities for scanning in azimuth and elevation, separately or in combination, are incorporated; these facilities permit the introduction of comprehensive monitoring procedures which may be carried out conveniently on a separate 'monitor' branch. Such monitoring, for example, would enable any significant changes in directions of wave arrival to be observed and the affected 'traffic' branches to be steered accordingly.

(3) POTENTIAL ADVANTAGES OF THE NEW FORM OF AERIAL SYSTEM

The basic form of the medusa type of aerial system described above permits a high degree of flexibility in both the design and the operation of installations to meet specific requirements.

Some of the potential advantages normally precluded with more conventional aerial systems are mentioned briefly.

The unit aerials can be positioned independently of each other, subject to the general requirements of small mutual impedance, an irregular array pattern and substantially uniform aerial density. This freedom in choosing aerial positions facilitates the accommodation of an array on a site of irregular shape and/or limited availability, e.g. enclosing streams, roads, gates or small buildings.

An important aspect of this freedom of choice in positioning aerials is that it enables the overall aperture of an array and the number of aerials (and therefore the average aerial-to-aerial spacing) to be chosen to suit the requirements of the particular installation being considered. In general, the optimum choice will be a compromise between economic considerations and the degree (dictated by operational requirements) to which the directivity characteristics of the aerial system are required to suit the relevant propagation conditions.[11]

When a large array can be justified, the virtual absence of limitations in the arrangement of the unit aerials offers the possibility of an array design determined solely by considerations of the nature of the signals being received. It has been shown that a radio wave after transmission via the ionosphere may be considered to consist of a steady specularly reflected component and a varying diffracted component of random amplitude and phase.[12–14] For typical propagation conditions the latter component predominates. Since, in general, the types of short-wave aerial in common use do not discriminate between the two components, the output from such an aerial arises mainly from the diffracted component of the received signal: it therefore displays the fading characteristics normally associated with short-wave reception. A medusa array, however, can be designed to discriminate against the diffracted component by using a large number of aerials, together with wide aerial-to-aerial spacings such that there is a significant spaced-aerial diversity effect between the aerials of the array. Under these conditions only those portions of the aerial voltages which are induced by the specularly reflected component of the wanted signal will be co-phased effectively in response to the phasing instructions: all the other constituents of the aerial voltage will remain randomly phased. The successful realization of this mode of operation would result in a substantial reduction of the effects of fading relative to those experienced with more conventional types of aerials.

The medusa type of aerial system may be extended to incorporate a 3-dimensional array by including unit aerials positioned at convenient heights above the ground. An array of this type may be desirable when a restricted beam-width in the vertical plane is required and when the available site area is limited.

By using a suitable type of unit aerial, e.g. a horizontally-polarized dipole, a steerable aerial system can be realized which is suitable for signals arriving at angles of elevation up to the zenith. Such a system would be of value in the reception of high-frequency signals from aircraft or earth satellites.

The method of phasing by the control of electronic switches offers a convenient means, applicable to branches separately, of effectively omitting individual aerials from an array when desired (by opening all the associated electronic switches independently of the phasing instructions). This facility permits modification of the directivity characteristic of the aerial system to meet operational requirements. For example, a reasonably constant beam-width over a wide frequency range could be achieved by omitting the outer aerials of an array when operating at the higher frequencies.

The use of a high-speed computer type of direction control favours the introduction of various 'automatic' features. For example, the outputs of a number of branches, associated with different circuits, could be examined sequentially and any necessary adjustments applied, preferably automatically, to the appropriate directions of steer. The output of each branch could be so examined at intervals of a few minutes, and the steering adjusted whenever necessary. Such adjustment would reduce the effects of any slow variations of the direction of arrival of the wanted signal which arise, for example, from movements of large-scale irregularities in the ionosphere,[15] and would maintain maximum signal level into a receiver energized from the system.

Another possible automatic feature is continuous signal-following, whereby a specially equipped branch of the system is locked on the wanted signal and follows continuously any variations of the direction of arrival. Such an arrangement may be used as a direction-finder by arranging for the continuous display of the automatically determined bearings and elevations. The system possesses the advantages discussed by Bain[6] of being of wide aperture and of providing information in respect of both azimuth and elevation. In this application the use of a large array would facilitate the measurement of low angles of elevation[16] and should also minimize site errors.[17]

For an installation in which only limited facilities are required, e.g. communication with a small number of distant stations, the provision of a high-speed computer direction-control with its full flexibility of operation may be unnecessary and/or uneconomic. In such a case the computer could be replaced by a memory device in which precalculated phasing instructions were stored, together with suitable read-out equipment. Alternatively, a centralized direction control could distribute phasing instructions to two or more medusa installations situated perhaps hundreds of miles apart, using, for example, standard v.f. telegraph channels.

A desirable feature of the medusa type of aerial system is the inherent protection against any catastrophic failure of the h.f. equipment. The system comprises a large number of parallel h.f. paths, namely aerial, amplifier, distribution, phasing and combining units, and the failure of any one is of limited consequence. The small number of basic types of unit is a further useful feature in that it simplifies both production and maintenance.

All the above considerations have been discussed in terms of a steerable aerial system for reception, but the majority of them apply equally to the problem of devising an equivalent system for transmission. There are additional considerations which are peculiar to the latter, e.g. broad-band power-handling amplifiers, which will not be pursued here.

(4) ACKNOWLEDGMENTS

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A FERRITE BOUNDARY-VALUE PROBLEM IN A RECTANGULAR WAVEGUIDE

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SUMMARY

Using a published formulation of the equations for the reflection of an electromagnetic wave from a transversely magnetized ferrite block in a rectangular waveguide, a solution in closed form for the integral equation involved is given. A detailed discussion of this solution in a certain range of the gyromagnetic parameter follows. Outside this range the published solutions are shown to be faulty, and the present solution is also unusable, for unknown reasons. A second-order solution is also given, and reasons are advanced for believing that quasi-static approximations are not the cause of the difficulties encountered.

LIST OF SYMBOLS

a = Width of rectangular waveguide.
A, B, C, D = Constants.
E = R.F. electric field.
F = Function associated with electric field at ferrite boundary.
H = Applied magnetostatic field.
K, M = Constants of the ferrite medium [see eqns. (4) and (5)].
$L = \frac{1}{\pi}\log\frac{K+M}{K-M}$
M_s = Saturation magnetization of the ferrite.
n = Integer describing nth mode.
p = Parameter given by eqn. (7).
q = A constant.
S = Residual effects of higher modes [eqn. (39)].
x, y, z = Rectangular co-ordinates, and also integration variables.
X = Interface reactance.
$Y = (1-y)/(1+y)$.
y_n = Integrals [see eqn. (30)].
α = Empirical damping parameter.
γ = Gyromagnetic ratio ($+2\cdot8$ Mc/s gauss).
$\gamma_n^{(1)}, \gamma_n^{(2)}$ = Attenuation constants of nth mode [eqn. (41)].
Δ_n = Parameter of nth mode [eqn. (40)].
ϵ = Permittivity of dielectric filling.
ϵ_f = Permittivity of ferrite.
κ, μ = Components of permeability tensor.
μ_0 = Permeability of free space.
$\mu_\perp$ = Effective permeability of transversely polarized ferrite.
θ, ϕ = Normalized variables.
ψ = Phase of $(M+K)/(M-K)$.
ω = Angular frequency.

(1) INTRODUCTION

We consider here a rectangular waveguide of width a filled with dielectric from $-\infty < z < 0$ and with ferrite from $0 < z < +\infty$. A transverse static magnetic field is applied over the ferrite region. On account of the gyrotropic properties of the ferrite, it is unable to support a mode with purely sinusoidal variation of magnetic field over the waveguide cross-section, and as a result, the boundary conditions at the interface cannot be met by assuming only the usual transmitted and reflected waves in the dominant mode; a complete series of modes is necessary and results in an infinite set of algebraic equations for their determination. No simple means is known for the resolution of these equations, and in a recent paper,* Sharpe and Heim avoid this difficulty by setting up an integral equation for the electrical field at the boundary, thus reducing the solution of the problem to the solution of this integral equation. [References to this paper and to its equations are denoted here by (A).] This solution is achieved by means of an iterative expansion obtained for both small and large values of a parameter which measures the strength of the gyrotropic effect. From the expression for the field a value can be calculated for the discontinuity reactance, jX, which appears at the dielectric-ferrite interface in the equivalent circuit for the arrangement. X is a measure of the energy stored in the higher-order modes at the junction. Among other approximations made is the common one of using the quasi-static form for the attenuation coefficients of the higher-order modes, all of which are assumed to be beyond cut-off.

(2) SHORTCOMINGS OF THE PUBLISHED SOLUTIONS

A number of difficulties are met in the above method of solution. The range of convergence of the expansions is not known, and there is an indication of divergence for intermediate values of the expansion parameter for which, in any case, the method fails to give a usable result. Secondly, the solution gives, not the electric field, but its differential. This may seem a trivial matter, but since the electric field must vanish at the edges of the guide, this requires that $\int_0^a E'(x)dx = 0$ and this equation is *not* automatically satisfied by the iterative solution. It turns out that the small-parameter expansion, but not the large-parameter one, satisfies this condition, the reason being that, for the former, in the process of arrangement of the equation prior to iteration, a result of Schmeidler was used which, in effect, introduces an arbitrary constant of the nature of an integration constant into the solution, and this constant is used to impose the necessary boundary condition. No such constant occurs in the large-parameter case, and the need for one is shown clearly by the fact that the dominant term in the solution involves $\sin \pi x/a$ which does not integrate to zero over the boundary, as required.

As a consequence, eqn. A(29) cannot be used in preference to eqn. A(12) for the reactance, which now (wrongly) turns out to be complex, even for a lossless medium. The solution for the large-parameter case must therefore be presumed inadequate and an alternative procedure must be sought.

Some of these difficulties, in particular those relating to the divergence for the intermediate case, might, it has been suggested, be a consequence of the inaccuracy inherent in the

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
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* SHARPE, C. B., and HEIM, D. S.: 'A Ferrite Boundary-Value Problem in a Rectangular Waveguide', *Transactions of the Institute of Radio Engineers*, 1958, MTT-6 p. 42.

quasi-static approximations used in the solution. However, as will be shown, the expansion parameter can take intermediate values in a range for which the quasi-static approximation is more than usually precise, and it is doubtful whether the cause is to be sought here.

(3) THE INTEGRAL EQUATION FOR THE FIELD

We shall use here the notation and expressions from (A). Our intention is to solve the equation

$$C \sin \phi = ME'(\phi) + \frac{jK}{\pi}\int_0^{\pi} \frac{E'(\theta) \sin \phi}{\cos \theta - \cos \phi} d\theta \quad . \quad . \quad (1)$$

and to calculate the discontinuity reactance X.

$$jX = (2/\pi) \int_0^{\pi} E'(\phi) \cos \phi d\phi \quad . \quad . \quad . \quad . \quad (2)$$

The first is eqn. A(18) and is arrived at as follows:

The field in the dielectric-filled part of the guide is expressed as an infinite sum of waveguide modes with undetermined coefficients. The field in the ferrite-filled guide is similarly expressed. On equating the tangential fields at the boundary, an equation for the coefficients results. These are then expressed, by Fourier analysis, in terms of the as yet undetermined field at the boundary, and by substituting for the coefficients, an integral equation involving the boundary field is obtained. Finally, by making the usual quasi-static approximation for the propagation coefficients of the higher modes, the summation involved can be expressed in closed form, leading to eqn. (1) after an integration by parts. The full details are given in the reference. Eqn. (2) results from integrating eqn. A(12) by parts and discarding the integrated term, since $E(0) = E(\pi) = 0$.

Eqn. (2) is preferred to eqn. A(29) which uses explicitly only the imaginary component of $E'(\phi)$, since the vanishing of the integration of the real part, which is to be expected for a loss-less medium, has to be confirmed rather than presumed. In any case, in a rigorous solution, the difference should not matter.

$E(\phi)$ is the electric field across the boundary, ϕ being a normalized co-ordinate given by $\phi = \pi x/a$. (The dash denotes differentiation with respect to the argument.) C is a constant given by

$$-C = 1 + KX \quad . \quad . \quad . \quad . \quad (3)$$

[not $1 - KX$ as erroneously shown in A(16)].

M and K express properties of the medium and are defined by

$$M = \frac{\pi}{a\omega} \frac{\kappa}{\mu^2 - \kappa^2} \quad . \quad . \quad . \quad . \quad . \quad (4)$$

$$K = \frac{\pi}{a\omega}\left(\frac{1}{\mu_0} + \frac{1}{\mu_\perp}\right) \quad . \quad . \quad . \quad . \quad . \quad (5)$$

μ and κ are the components of the permeability tensor

$$\begin{bmatrix} \mu & 0 & j\kappa \\ 0 & \mu_0 & 0 \\ -j\kappa & 0 & \mu \end{bmatrix}$$

$\mu_\perp$ is the effective permeability of the transversely polarized ferrite and is given in terms of μ and κ by

$$\mu_\perp = (\mu^2 - \kappa^2)/\mu \quad . \quad . \quad . \quad . \quad . \quad (6)$$

It is seen that M, which vanishes with κ, depends on the gyrotropic properties of the ferrite. The quantity M/K is the expansion parameter referred to previously, and the difficulties met seem to be associated with values of M/K near ± 1. In particular, it will be shown later that the expression

$$p = \frac{1}{2\pi j} \log \frac{M + K}{M - K} \quad . \quad . \quad . \quad . \quad . \quad (7)$$

is an important quantity in the analysis, and it is necessary to examine carefully the changes associated with the singularities of this expression.

(4) THE VARIATION OF M AND K

It is possible to express μ and κ in terms of the ferrite saturation magnetization, M_s, the static applied field, H, and the gyromagnetic ratio γ (+ 2·8 Mc/s per gauss). The equations given in (A) can be combined to give

$$\frac{\mu \pm \kappa}{\mu_0} = 1 + \frac{\gamma M_s}{\gamma H \pm \omega} \quad . \quad . \quad . \quad . \quad . \quad (8)$$

This equation represents the form for the loss-free case. In practice the medium must exhibit some loss, however slight, and this is met by the addition of an empirical damping term $j\alpha$ in the denominator of eqn. (8):

$$\frac{\mu \pm \kappa}{\mu_0} = 1 + \frac{\gamma M_s}{\gamma H \pm \omega + j\alpha} \quad . \quad . \quad . \quad . \quad (9)$$

Note that α takes the same sign as H and M_s. The addition of this term permits an adequate examination of the phase at resonance of the various quantities concerned. The loss-free case can then be recovered by setting $\alpha = 0$ at the end of the analysis. From eqns. (4), (5), (6) and (9) we get

$$\frac{M + K}{M - K} = -\frac{\gamma(H + M_s) + \omega + j\alpha}{\gamma(H + \frac{1}{2}M_s) + \omega + j\alpha}\,\frac{\gamma(H + \frac{1}{2}M_s) - \omega + j\alpha}{\gamma(H + M_s) - \omega + j\alpha} \quad . \quad . \quad . \quad . \quad (10)$$

Consider first the case of negative applied field (both α and M_s will be negative). Then the quotient on the extreme right of eqn. (10) is essentially real, as is also the quotient to its left, except when in the neighbourhood of $\omega = -\gamma(H + \frac{1}{2}M_s)$ or $\omega = -\gamma(H + M_s)$. As the field is decreased from zero the phase varies from -1 to $+j$, at which point $\gamma(H + M_s) = -\omega$ (α is assumed small but non-zero). As the field still further decreases, the phase runs round to nearly zero and then increases again to $+j$ at $\gamma(H + \frac{1}{2}M_s) = -\omega$. Beyond this value the phase increases to nearly -1. Thus, $0 <$ phase $(M + K)/(M - K) < \pi$ for negative applied field. For positive values of H, α and M_s the first quotient in eqn. (10) is substantially real and positive, the second, together with the initial minus sign, giving $-j$ at $\gamma(H + M_s) = \omega$. The phase runs round to nearly 2π and returns to $-j$ at $\gamma(H + \frac{1}{2}M_s) = \omega$. Thereafter it approaches -1 again.

Thus, for positive or negative field we have

$$0 < \text{phase}\,(M + K)/(M - K) < 2\pi \quad . \quad . \quad (11)$$

Hence, $\log (M + K)/(M - K)$ involves an angle $j\psi$ where $0 < \psi < 2\pi$ and the real part of the parameter p in eqn. (7) accordingly involves the quantity $\psi/2\pi$ which lies between zero and one. Since zero phase angle is approached but not actually attained, we have

$$0 < \mathscr{R}p < 1 \quad . \quad . \quad . \quad . \quad . \quad (12)$$

and the real part of p is always positive and less than unity. This result is extremely important for what follows. The approach to zero phase is the crucial point here, and it is worth

re-writing eqn. (10) in a form which shows explicitly that zero phase itself is not actually attained with a non-zero value of α:

$$\frac{M+K}{M-K} = -\left[1 + \frac{\frac{1}{2}\gamma M_s}{\gamma(H+\frac{1}{2}M_s)+\omega+j\alpha}\right]\left[1 - \frac{\frac{1}{2}\gamma M_s}{\gamma(H+M_s)-\omega+j\alpha}\right] \quad . \quad . \quad (13)$$

If we take the case of negative field, and set $-\gamma(H+\frac{1}{2}M_s) = \omega + \delta$, the first factor, together with the minus sign, is $-1 + \frac{1}{2}\gamma M_s/(\delta - j\alpha)$ and always has a positive phase angle (note that α is negative for negative field). The second factor, whose effect is negligible anyway, also has a positive phase angle. Thus zero phase is never actually attained. When the field and magnetization are positive, we similarly put $\gamma(H+M_s) = \omega + \delta$. The second factor is $1 + (-\frac{1}{2}\gamma M_s)/(\delta + j\alpha)$ which gives a negative imaginary part in conjunction with the initial minus sign in eqn. (13). The first factor also has a negative imaginary part in this region. Thus a phase of 2π is never actually attained. This confirms the inequality sign in condition (11).

We can now examine the variation of $\mu_\perp$, the effective permeability for propagation in transversely magnetized ferrite. From eqns. (6) and (9) we get

$$\frac{\mu_\perp}{\mu_0} = \frac{[\gamma(H+M_s)+\omega-j\alpha][\gamma(H+M_s)-\omega-j\alpha]}{(\gamma H - j\alpha)[\gamma(H+M_s)-j\alpha]-\omega^2} \quad . \quad (14)$$

For positive fields, and in the absence of a damping term, $\mu_\perp$ goes through zero at $\gamma(H+M_s) = \omega$, and then decreases and passes through $-\infty$ to $+\infty$ as $\gamma^2 H(H+M_s) = \omega^2$.

The singularity occurring in eqn. (7) at $M = K$ is seen from eqn. (10) to take place at $\gamma(H+M_s) = \omega$, i.e. when $\mu_\perp$ goes through zero. Now the quasi-static approximation for the evanescent modes involves putting $n\pi/a$ for $\sqrt{[(n\pi/a)^2 - \omega^2\epsilon_f\mu_\perp]}$. That is, just in the region where the first singularity occurs, the quasi-static approximation is seen to hold the best, since $\mu_\perp$ is zero there. It is therefore difficult to see that the problems associated with the present solution can have been basically affected by the use of the quasi-static form. It is apparent, however, that this form is unusable for large values of $\mu_\perp$, negative or positive, and the solution must break down at $\gamma^2 H(H+M_s) = \omega^2$. The singularity at $M = -K$ in eqn. (10), however, occurs at $\gamma(H+\frac{1}{2}M_s) = \omega$ which is slightly earlier (for an increasing H) than the value for which $\mu_\perp$ becomes infinite.

(5) THE SOLUTION OF THE INTEGRAL EQUATION

In eqn. (1) let $\cos\theta = x$, $\cos\phi = y$, and $E'(\phi) = \sin\phi F(y)$. The integral equation becomes

$$C = MF(y) + \frac{jK}{\pi}\int_{-1}^{1}\frac{F(x)}{x-y}dx \quad . \quad . \quad . \quad (15)$$

(The principal value of the integral on the right is to be understood.) Eqn. (2) for the reactance becomes

$$jX = \frac{2}{\pi}\int_{-1}^{1} yF(y)dy \quad . \quad . \quad . \quad . \quad (16)$$

whilst the condition $E(\pi) = E(0) = 0$ becomes

$$\int_{-1}^{1} F(y)dy = 0 \quad . \quad . \quad . \quad . \quad . \quad (17)$$

An indication of the form of solution which might exist for eqn. (15) is given by eqns. A(33) and A(34) which involve powers of $\log(1-y)/(1+y)$. Such a structure could arise from the expansion in a power series in a parameter p of some such form as $[(1-y)/(1+y)]^p$, and it was in the investigation of this form that a solution to eqn. (15) was discovered. However, this solution does not satisfy eqn. (17), and, as mentioned earlier, something in the nature of an integration constant is also required. This would be a solution of the homogeneous equation

$$0 = MF(y) + \frac{jK}{\pi}\int_{-1}^{1}\frac{F(x)}{x-y}dx \quad . \quad . \quad . \quad (18)$$

From an examination of the non-gyromagnetic case ($M = 0$), for which p turns out to be one half, and the form taken by Schmeidler's solution in that case, a term $B/(1-y^2)^{\frac{1}{2}}$ is indicated. And if this is generalized to $[(1-y)/(1+y)]^p B/(1-y)$ where B is an arbitrary constant, it can be verified that eqn. (18) is indeed satisfied. Hence we shall examine the properties of the function

$$F(x) = A\left(\frac{1-x}{1+x}\right)^p\left(1 + \frac{2B}{1-x}\right) \quad . \quad . \quad . \quad (19)$$

A, B and p are constants at our disposal. We form the principal value of the integral $I = \int_{-1}^{1}\frac{F(x)}{x-y}dx$:

$$I = A\int_{-1}^{1}\left(\frac{1-x}{1+x}\right)^p\left(1+\frac{2B}{1-x}\right)\frac{dx}{x-y} \quad . \quad . \quad (20)$$

Let us change the variable, putting $(1-x)/(1+x) = z$, and write $(1-y)/(1+y) = Y$ for short. After a little manipulation, eqn. (20) becomes

$$I = A\int_0^\infty z^p\left[\frac{1}{1+z} + \frac{1}{Y-z} + \frac{2B}{1-y}\left(\frac{1}{z} + \frac{1}{Y-z}\right)\right]dz \quad . \quad (21)$$

For convergence at z near zero we need $\mathscr{R}p > 0$, whilst for convergence at $z = \infty$ we need $\mathscr{R}p < 1$. We assume, therefore, that the parameter p is constrained within these limits.

Now $\int_0^\infty \frac{z^{q-1}}{1+z}dz = \pi \operatorname{cosec} q\pi$ and $\int_0^\infty \frac{z^{q-1}}{1-z}dz = \pi \cot q\pi$ are two well-known integrals, valid for $0 < \mathscr{R}q < 1$. Bearing in mind the ranges of p and q, we seek to integrate eqn. (21), using these results.

If we put Yz for z in $\int_0^\infty \frac{z^{q-1}}{Y-z}dz$ we get $Y^{q-1}\int_0^\infty \frac{z^{q-1}}{1-z}dz$.

Hence, $\int_0^\infty z^{q-1}\left(\frac{1}{1+z} + \frac{1}{Y-z}\right)dz =$

$$\int_0^\infty z^{q-1}\left(\frac{1}{1+z} + \frac{Y^{q-1}}{1-z}\right)dz = \frac{\pi}{\sin\pi q}[1 + Y^{q-1}\cos q\pi]$$

Now this equation is proved only for $0 < \mathscr{R}q < 1$, but both sides are analytic in the extended range $0 < \mathscr{R}q < 2$, since the integral on the left is convergent in this range, whilst the expression on the right does not exhibit any singularity, the point $q = 1$ being an ordinary point since $1 + Y^{q-1}\cos q\pi$ goes to zero linearly there. The equation is therefore valid in the extended range. In particular, if we put $q = 1 + p$ we get

$$\int_0^\infty z^p\left(\frac{1}{1+z} + \frac{1}{Y-z}\right)dz = \frac{\pi}{\sin p\pi}(Y^p\cos p\pi - 1) \quad . \quad (22)$$

which is certainly valid over the smaller range $0 < \mathscr{R}p < 1$.

For the remaining terms in eqn. (21) we can put

$$z^p\left(\frac{1}{z}+\frac{1}{Y-z}\right)dz = z^p\left(\frac{1}{z}+\frac{1}{1-z}+\frac{1}{Y-z}-\frac{1}{1-z}\right)$$
$$=\frac{z^{p-1}}{1-z}+z^p\left(\frac{1}{Y-z}-\frac{1}{1-z}\right).$$

The first of these integrates to $\pi \cot p\pi$, whilst the second pair, by a process almost identical to that which led to eqn. (22) gives

$$\int_0^\infty z^p\left(\frac{1}{Y-z}-\frac{1}{1-z}\right)dz = \pi \cot p\pi(Y^p - 1) \quad . \quad (23)$$

again valid for $0 < \mathscr{R}p < 1$. Collecting terms we find

$$I = -A\pi \operatorname{cosec} p\pi + A\pi \cot p\pi Y^p\left(1+\frac{2B}{1-y}\right)$$

Hence,

$$\int_{-1}^{1}\frac{F(x)}{x-y}dx = -A\pi \operatorname{cosec} p\pi + \pi \cot p\pi\, F(y) \quad . \quad (24)$$

Comparing this with eqn. (15), we see that the two equations will be the same if we take

$$\left.\begin{aligned} A &= (jC/K)\sin p\pi \\ \text{and} \qquad \tan p\pi &= (-jK/M) \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad (25)$$

From the second of these, using the formula

$$\operatorname{arc}\tan x = \frac{1}{2j}\log\frac{1+jx}{1-jx}$$

we get
$$p = \frac{1}{2\pi j}\log\frac{M+K}{M-K} \quad . \quad . \quad . \quad . \quad (26)$$

which is the same as eqn. (7), whose properties have already been discussed. In particular, from expression (12) we see that the real part of p lies in the range necessary to validate the preceding analysis. Eqn. (26) is the form suitable for large values of M/K. For small values we write

$$(M+K)/(M-K) = \varepsilon^{j\pi}(K+M)/(K-M),$$

and eqn. (26) becomes

$$p = \frac{1}{2}+\frac{1}{2\pi j}\log\frac{K+M}{K-M} \quad . \quad . \quad . \quad . \quad (27)$$

This confirms an earlier remark that $p = \frac{1}{2}$ when $M = 0$. It remains to determine B so that eqn. (17) shall be satisfied. Inserting eqn. (19) into eqn. (17), again making the substitution $(1-x)/(1+x) = z$, gives

$$\int_0^\infty z^p\left[\frac{1}{(1+z)^2}+\frac{B}{z(1+z)}\right]dz = 0 \quad . \quad . \quad . \quad (28)$$

This and the subsequent analysis involves the integrals

$$y_n = \int_0^\infty z^p(1+z)^{-n}dz.$$

By differentiating $z^{p+1}(1+z)^{1-n}$, rearranging and integrating, we get the recurrence formula

$$y_{n+1} = \left(\frac{n-1-p}{n}\right)y_n \quad . \quad . \quad . \quad . \quad (29)$$

Since $y_1 = -\pi \operatorname{cosec} \pi p$, we get

$$\left.\begin{aligned} y_2 &= \pi p \operatorname{cosec} \pi p \\ y_3 &= \tfrac{1}{2}\pi p(1-p)\operatorname{cosec} \pi p, \text{ etc.} \end{aligned}\right\} \quad . \quad . \quad . \quad (30)$$

With the exception of y_1, the range of validity, which is determined by the convergence of the integral at its limits, includes the domain $0 < \mathscr{R}p < 1$. Returning to eqn. (28) we get $\pi \operatorname{cosec} \pi p\,[p + B] = 0$. Hence

$$B = -p \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (31)$$

We can now use eqn. (16) to determine the interface reactance, X.

$$jX = (jC/K)\sin(p\pi)\frac{2}{\pi}\int_{-1}^{1} y\left(\frac{1-y}{1+y}\right)^p\left(1-\frac{2p}{1-y}\right)dy \quad . \quad (32)$$

The substitution $(1-y)/(1+y) = z$ reduces this integral to a form similar to eqn. (28), the integrations involving y_2 and y_3 of eqn. (30). After a little simplification this gives

$$X = -\frac{4C}{K}p(1-p) \quad . \quad . \quad . \quad . \quad (33)$$

If we use eqn. (27) for p this becomes

$$X = -\frac{C}{\pi^2 K}\left(\pi^2+\log^2\frac{K+M}{K-M}\right) \quad . \quad . \quad . \quad (34)$$

Combining this with eqn. (3), which gives C in terms of X, we get

$$X = \frac{-1}{K}\left[1+\pi^2/\log^2\left(\frac{K+M}{K-M}\right)\right] \quad . \quad . \quad . \quad (35)$$

This equation compares with A(30). If we take $M \ll K$ and expand $\log (K+M)/(K-M)$ as $2M/K$, eqn. (35) gives

$$X \simeq \frac{-1}{K}\left[1+\frac{\pi^2}{4}\left(\frac{K}{M}\right)^2\right] \quad . \quad . \quad . \quad . \quad (36)$$

which is the value given in A(30) for $|M/K| < 1$.

Accordingly, there is full agreement between the present analysis and (A) for the case $|M/K| < 1$. As explained earlier, (A) is faulty in the rest of the range because of the non-vanishing of $E(\pi)$, and agreement is not to be expected. On the other hand, the present analysis makes no distinction between the two ranges, and with the correct value for the logarithm, eqn. (35) should continue to hold. It is therefore extremely surprising to find that it gives complex values of X in the range $|M/K| > 1$. To be more explicit, the formula is

$$X = -\frac{1}{K}\left[1+\left(\frac{\pi}{-j\pi+\log\dfrac{M+K}{M-K}}\right)^2\right] \quad . \quad (37)$$

and since jX is really the interface impedance, eqn. (37) would appear to indicate the presence of a resistive part. It is apparent, however, that jX must be imaginary for conservation of power [eqn. (37) would even indicate a negative resistance over part of the range of the variables] and the conclusion must be that at some stage the mathematical analysis is at fault, though at what point is not clear. A number of possible causes have been investigated, including convergence of the integrals concerned, the multivalued nature of the logarithm and the derivation of eqn. (1) itself. No solution to the difficulty has appeared, but it is clear that at some stage an unnoticed restriction on the permissible mathematical operations must have entered. Until this difficulty has been cleared up the range of validity of eqn. (35) must be limited to $|M/K| < 1$, though the source of this restriction is at present unknown.

(6) A SECOND-ORDER APPROXIMATION

The solution given in eqn. (37) is a quasi-static one, dependent on approximations made for the attenuation coefficients of the

higher-order modes. This type of approximation is least valid for the second-order mode, and we shall show here how a solution can be obtained which uses the correct form for the attenuation constant of this mode. In principle the method could be extended to include any finite number of higher-order modes, though the complexity of the analysis becomes prohibitive when the number is at all large.

The value of extending the analysis to a higher-order solution is threefold. First, the correction term gives an indication of the accuracy attained by the quasi-static solution; secondly, when the second-order mode propagates, the solution shows what fraction of the power is conveyed in this mode; and thirdly, the nature of the solution indicates quite clearly that the difficulties met for $|M/K| > 1$ cannot be removed by considering a finite number of high-order terms, but that these difficulties must be taken as intrinsic to the problem.

Eqn. A(13) gives the rigorous expression for the integral equation for the field. It can be put in the form

$$C \sin \phi = ME'(\phi) + j(K/\pi)\int_0^\pi \frac{E'(\theta)\sin\phi}{\cos\theta - \cos\phi}d\theta + S \quad . \quad (38)$$

where

$$S = j(2K/\pi)\sum_2^\infty \Delta_n \int_0^\pi E'(\theta)\sin n\phi \cos n\theta d\theta \quad . \quad . \quad (39)$$

This compares with eqn. (1). It differs by the additional term, S, which involves the quantities Δ_n which measure the departure of the higher-order mode attenuation coefficients from their quasi-static values. More specifically,

$$\Delta_n = \frac{\mu_\perp\left[\frac{a}{\pi}\gamma_n^{(1)} - n\right] + \mu_0\left[\frac{a}{\pi}\gamma_n^{(2)} - n\right]}{n(\mu_0 + \mu_\perp)} \quad . \quad . \quad (40)$$

where

$$\left.\begin{aligned}\gamma_n^{(1)} &= \sqrt{[(n\pi/a)^2 - \omega^2\epsilon\mu_0]}\\ \gamma_n^{(2)} &= \sqrt{[(n\pi/a)^2 - \omega^2\epsilon_f\mu_\perp]}\end{aligned}\right\} \quad . \quad . \quad . \quad (41)$$

and ϵ and ϵ_f are, respectively, the permittivity of the dielectric and ferrite in the two parts of the guide. It is seen that, so long as $\omega^2\epsilon\mu_0$ and $\omega^2\epsilon_f\mu_\perp$ are each sufficiently less than $(n\pi/a)^2$, the approximation $\gamma_n \simeq (n\pi/a)$ gives $\Delta_n \simeq 0$ and S may be ignored. The term for which this approximation is least sufficient is that for which n is smallest, i.e. $n = 2$. Two separate expressions, $\gamma_n^{(1)}$ and $\gamma_n^{(2)}$, are involved and the approximation may be much worse for one than for the other. In fact, it is possible for $\gamma_n^{(2)}$ to be imaginary, representing power flow in the second-order mode in the ferrite, whilst $\gamma_n^{(1)}$ may still refer to a guide beyond cut-off. Thus the two parts involved in eqn. (40) need not be of equal importance, but it is as easy to consider them both together as to deal with one only, and no separation will therefore be made.

We retain the $n = 2$ term only in eqn. (39), and write

$$\begin{aligned}S &= j(2K/\pi)\Delta_2\int_0^\pi E'(\theta)\sin 2\phi\cos 2\theta d\theta\\ &= j(4K/\pi)\Delta_2(y\sin\phi)\int_{-1}^1 (2x^2-1)F(x)dx \quad . \quad (42)\end{aligned}$$

on making the changes of variable which led to eqn. (15).

Now the integral in eqn. (42) is a constant (independent of y or ϕ). If we define a new constant, D, by

$$-j(4K/\pi)\Delta_2\int_{-1}^1 (2x^2-1)F(x)dx = CD \quad . \quad . \quad (43)$$

then the form taken by eqn. (38) is

$$C(1 + Dy) = MF(y) + (jK/\pi)\int_{-1}^1 \frac{F(x)}{x-y}dx \quad . \quad . \quad (44)$$

and this differs from eqn. (19) only by the term in Dy.

The addition of a linear term to eqn. (19) enables an equation of this form to be solved, and it is readily verified that a solution of eqn. (44) is

$$F(x) = (jC/K)\sin p\pi\left(\frac{1-x}{1+x}\right)^p\left(1 + 2Dp + Dx + \frac{2B}{1-x}\right) \quad . \quad . \quad . \quad . \quad (45)$$

where B has to be determined by the condition (17).

The method of integration used previously enables eqn. (45) to be integrated and the value of B to satisfy condition (17) is

$$B = -p(1 + Dp) \quad . \quad . \quad . \quad . \quad . \quad (46)$$

The value of the interface reactance is now obtained from eqn. (16) through the integration of $\int_{-1}^1 yF(y)dy$. After some simplification it is found that

$$jX = -4(jC/K)p(1-p)\left[1 + \frac{D(2p-1)}{3}\right] \quad . \quad (47)$$

It remains to determine D from eqn. (43), and this involves the evaluation of $\int_{-1}^1 (2x^2-1)F(x)dx$. The previous methods are used, giving $(j2C\pi/3K)p(1-p)[4(2p-1) - 6Dp(1-p)]$ for the value of this integral. Inserting this in eqn. (43) gives D, and if C and D are now eliminated between eqns. (47) and (3), an equation for X results which can be solved to give

$$X = -\frac{1}{K}\left[1 + \frac{1}{L^2}\frac{1 + \Delta_2(1+L^2)^2}{1 + \Delta_2(1+L^2)^2/9}\right] \quad . \quad . \quad (48)$$

where $L = \frac{1}{\pi}\log\frac{K+M}{K-M}$ and Δ_2 is given by putting $n = 2$ in eqn. (40). This solution reduces to eqn. (35) if Δ_2 is put equal to zero, and it is concluded that the presence of this additional term does not alter in any fundamental way the solution of the relevant equations. In particular, the limitations discussed earlier which led to eqn. (35) being restricted to the region $|M/K| < 1$ must be presumed to hold for eqn. (48) too.

(7) CONCLUSIONS

The method of solution proposed is adequate to treat the problem in the region $|M/K| < 1$ but further investigation is required to discover the reason for this restriction. When this has been done it should be possible to solve the problem in the remainder of the range.

(8) ACKNOWLEDGMENTS

The author wishes to thank several of his colleagues for valuable discussions on a number of aspects of this problem. Thanks are also due to Standard Telecommunication Laboratories, Ltd., for facilities to prepare, and permission to publish, the paper.

DISCUSSION ON

'THE PERFORMANCE OF DIRECTIVE AERIALS IN COMPLEX U.H.F. FIELDS'*

Messrs. P. Knight and **H. Page** (*communicated*): The complex field distribution encountered within the service area of a u.h.f. transmitter arises from the reception of a number of components arriving from different directions, as the result of natural or man-made irregularities of the terrain. Relative to the direct component—that arriving from the direction of the transmitter—the other components are delayed in time and may cause distortion in a television service. A receiving aerial should reject all the signals with long time delays and preferably also those reflected by nearby objects, since these signals, while not causing appreciable distortion, can oppose the direct signal and give rise to a low receiver input voltage. The field due to the direct signal alone, even though it be influenced by irregularities in the direction of the transmitter, will be plane and undistorted; for this component alone the aerial will therefore give its full gain. On the other hand, a simple dipole receives both wanted and unwanted components; it is therefore advantageous if the 'apparent gain' of a directional aerial differs from the gain in a plane-wave field, whether it be higher or lower, as this means that any distortion present has been eliminated. The authors' statements, which refer specifically to the band allocated for television broadcasting, could be construed as implying that a low apparent gain is a disadvantage.

This raises the question as to the best method of measuring the field strength in carrying out a survey for a television service. Since we are interested in receiving an undistorted signal, we should measure only the direct component; however, if there is a predominant delayed component which is stronger than that arriving from the direction of the transmitter, we should regard this as the component to be measured. It therefore seems desirable to carry out measurements using not a half-wave dipole but a highly directional aerial oriented to give maximum pick-up. We suggest, therefore, that the term 'apparent gain' is a misleading concept in the context of the paper. It does, however, appear to be useful in calculating the pick-up in the case of, say, a scatter-transmission service, in which the wanted signal is not necessarily a plane wave.

Dr. J. A. Saxton and **Mr. B. N. Harden** (*in reply*): The views expressed by Messrs. Knight and Page differ more in emphasis than in principle from those given in our paper. We were careful to point out that, well inside the service area, the most important attribute of a directional aerial may be its ability to exclude delayed signals. Nevertheless, if satisfactory coverage is to be obtained with a u.h.f. transmitter, it is essential that the most should be made of fringe-area signals, in which context the gain of the aerial is probably the most important factor.

It is doubtful if in practice it will always be possible to use receiving aerials with sufficient directivity to exclude all but one signal: apart from this reservation we agree that the method of making field-strength surveys suggested by Messrs. Knight and Page is probably the best one. A comparison is, in fact, given in the paper of the signal obtained with the aerial oriented for reception along the direct path from the transmitter and the maximum signal, whatever its direction of arrival.

In aerial gain measurements some standard of reference must be used, and the results obtained are, in effect, always 'apparent gains', which may differ according to the circumstances of the observations. We chose a dipole as standard: admittedly it may have limitations for this purpose, but it is still widely used as such. Some other (more directive) aerial could have been used as a standard; but it is clear from the shapes of the curves in Figs. 1 and 2 that the overall statistical appraisal of the relative performances of the various aerials (from the gain point of view) would not have been materially changed.

* SAXTON, J. A., and HARDEN, B. N.: Paper No. 2807 E, May, 1959 (see 106 B, p. 315).

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ABBREVIATIONS

(P)—Address, lecture or paper.
(p)—Subject dealt with in a paper or address.
(D)—Discussion on a paper.
(A)—Abstract of paper or address.

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